T. Detert, and A. Fernekeß, ”Joint MSE Channel Estimation within the Current GSM/EDGE Standard,” in *Proc. of IEEE Vehicular Technology Conference (VTC-Fall)*, Los Angeles, USA, Sept. 2004, pp. 3822-3827

Joint MSE Channel Estimation within the Current GSM/EDGE Standard

Thorben Detert
Institut für Nachrichtentechnik
Technische Universität Darmstadt
Merckstrasse 25, D-64283 Darmstadt, Germany
Email: t.detert@nt.tu-darmstadt.de

Andreas Fernekeß
Panasonic European Laboratories GmbH
Monzastraße 4c
D-63225 Langen, Germany

***Abstract*— In this paper, a joint mean square error (JMSE) approach is taken in order to perform channel estimation in a frequency selective fading environment in the downlink. It is intended to provide the basis for joint sequence estimation, such as joint delayed decision feedback sequence estimation (JDDFSE), in the mobile terminal. Up to three quasi-synchronous users are considered, one user of interest and two co-channel interferers. The simulation results indicate that this technique is suited for practical systems, such as the Global System for Mobile Communications (GSM) and its 8-PSK add-on, Enhanced Data Rates for Global Evolution (EDGE), if the base stations are synchronized, resulting in a delay of only several symbols at the mobile station. Simulation results in mixed GSM/EDGE networks are shown.**

I. Introduction

Co-channel interference cancellation has always been a major topic in mobile communication systems. When code division multiple access (CDMA) is used, co-channel interference is mainly caused by intracell interference due to imperfect cross-correlation properties of the spreading sequences. The co-channel interference present within time division multiple access (TDMA) systems stems mainly from intercell interferers. It is mitigated by having a frequency reuse factor larger than one, so the total number of frequency bands is shared amongst the cells, and the co-channel interferers lose their strength by pathloss.

Currently, a lot of research is done in order to allow for a smaller frequency reuse factor, thus increasing the capacity of TDMA systems, by cancelling the co-channel interference [1], [2], [3], [4], [5]. Currently, the topic is being treated in standardization to define new restrictions on GSM and EDGE air interface, see e.g. [6]. ARP (Advanced Receiver Performance) is the term coined by the third generation partnership project (3GPP) for this research area of huge practical interest. Probability considerations on the number of occurring co-channel interferers and their strengths can be found in [7], [8] and [9].

The idea is to carry out interference cancellation (IC), as it is done in CDMA systems. These approaches suffer from limited knowledge of the interfering users' channels. The only study for estimating the co-channel impulse responses jointly by an MSE approach seems to be [10], where multiple users arrive at the base station, thus in the uplink. Their relative delays are due to coarse symbol synchronization. In this paper, the pioneering work in terms of the signal model and estimator presented in [10] is applied in the downlink and refined to

- include the rotation and derotation techniques used in the current GSM/EDGE standard
- allow for co-channel signals with different modulation formats.

The JMSE technique implemented is dependent on a synchronization of the base stations, resulting in shift between the training sequences of co-channel users of only a few symbols. With Laurent's decomposition [11] and a derotation receiver [12], binary continuous phase modulation (CPM) signals of modulation index 0.5, such as Gaussian minimum shift keying (GMSK), can be processed. This approach is taken here.

The paper is organized as follows. In section II, the signal model suggested in [10] is repeated and refined to the GSM/EDGE signal model in the downlink, followed by the modified estimator in section III. In section IV, interference scenarios as suggested by 3GPP TSG GERAN[1] are given. The performance results are discussed in section V. Finally, some conclusions are given in section VI.

II. The Quasi-Synchronous Transmission Model

The signal model suggested in [10] has to be refined to take into account the rotation and derotation used within GSM/EDGE.

A. GSM GMSK and 8-PSK signal models

The bandpass signal of a binary GMSK modulator can be achieved by a simple filtering operation of the transmitted data pulses by a pulse shaping filter, whose output signal is applied to a voltage controlled oscillator with center frequency f_c, see Fig. 1 for user u, $u = 1, 2, \ldots, U$.

The equivalent baseband signal $s_u(t)$ of $s_{BP,u}(t)$ can be represented as

$$s_u(t) = \exp\left(\frac{j\pi}{2}\sum_{n=-\infty}^{\nu} c_u(n)\int_0^t g(\tau - kT)d\tau + j\theta_0\right), \quad \nu T \le t < (\nu+1)T. \tag{1}$$

[1]Technical Specification Group (TSG) GSM/EDGE Radio Access Network (GERAN).

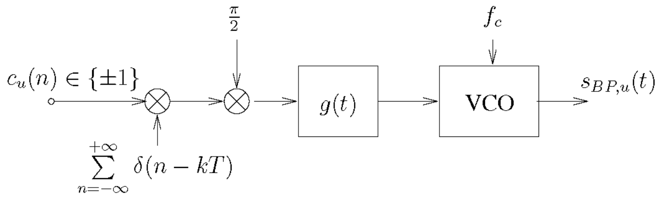


Fig. 1. The GMSK modulator for the u-th user.

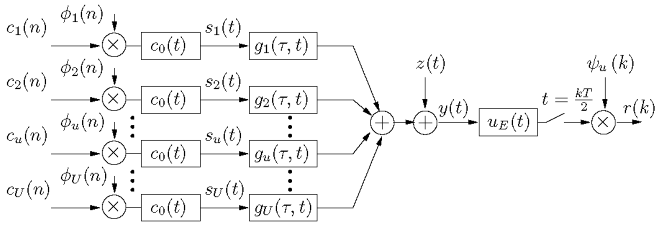


Fig. 2. Co-channel signal transmission.

$c_u(n)$	is the data sequence of the u-th user derived from original (binary) data sequence $x_u(n)$ by differential precoding: $c_u(n) = x_u(n) \cdot x_u(n-1)$; with: $c_u(n), x_u(n) \in \{\pm 1\}$
T	is the symbol interval
θ_0	is a constant starting phase

(2)

and the filter impulse response $g(t)$ given by

$$g(t) = \frac{1}{2}\left[\operatorname{erf}\left(\gamma\left(\frac{t}{T}+\frac{1}{2}\right)\right) - \operatorname{erf}\left(\gamma\left(\frac{t}{T}-\frac{1}{2}\right)\right)\right]$$

with $\operatorname{erf}(\cdot)$ the error function and

$$\gamma = \sqrt{\frac{2}{\ln 2}} \cdot \pi \cdot B \cdot T. \tag{3}$$

B is the $3dB$ cutoff frequency. We choose $B \cdot T = 0.3$ as in the GSM standard.
The constant phase θ_0 is set to zero without loss of generality. According to [13], 99.7236% of the energy in $s(t)$ is in the pulse $c_0(t)$ of the Laurent decomposition, thus, the GMSK signal can be approximated by a pulse amplitude modulated (PAM) signal with the pulse $c_0(t)$ as the pulse shaper only.
Whereas the data symbols $c_u(n)$ are taken from the alphabet -1 and $+1$ in the case of GMSK, they are $e^{j2\pi\frac{l}{8}}$ with $l \in [0, 1, \ldots, 7]$ for 8-PSK. The same $c_0(t)$ transmit filter is utilized. In the case of GMSK, the symbols are rotated with $\phi_u(n) = j^n$, for 8-PSK with $\phi_u(n) = e^{j3\pi/8 \cdot n}$.

B. The transmitted signals as seen by the receiver

As most energy is covered by the pulse $c_0(t)$, in the derivation of the channel estimator, a PAM transmission following Laurent's decomposition is assumed, although in truth the transmitted signal stems from a binary CPM modulator. The signal is transmitted in bursts. Fig. 2 shows the underlying signal model of the asynchronous transmission. We stick to the signal model and, partly, to the notation given in [10]. With differential precoding and Laurent decomposition as assumed above, the transmitted signal $s_u(t)$ is approximated by

$$s_u(t) = V_u \sum_n c_u(n)\phi_u(n)u(t - nT - \tau_u). \tag{4}$$

V_u is the u-th user's amplitude and can be computed from the average power P_u as $V_u = \sqrt{P_u}$. τ_u denotes the delay of the u-th user, covering both delayed transmission and the minimum delay of the impulse response seen by the path with shortest distance. The data symbols $c_u(n)$ are -1 and $+1$ when GMSK modulation is used, and $e^{j2\pi\frac{l}{8}}$ with $l \in [0, 1, \ldots, 7]$ for 8-PSK modulation. The symbols are assumed to be rotated by $\phi_u(n) = j^n$ with GMSK transmitters and $\phi_u(n) = e^{j3\pi/8 \cdot n}$ with 8-PSK transmitters. The channel impulse responses follow

$$g_u(\tau, t) = \sum_n \alpha_n(t) e^{-j2\pi f_c \tau_n(t)} \delta(\tau - \tau_n). \tag{5}$$

The individual channel impulse responses $g_u(\tau, t)$ are assumed to be statistically independent. A white Gaussian noise process $Z(t)$ with two sided power spectral density N_0 is added. The receive filter is a root-raised cosine filter. A sufficient statistics for maximum likelihood sequence estimation is provided only in extreme cases, however, the loss is negligible [14]. The deterministic cross-correlation function (CCF) of the transmit and receive filters is given by $x(\alpha) = \int c_0(t)u_R(\alpha - t)dt$. After the receive filter, a derotation with $\psi_u(t) = (-j)^{t/T}$ or $\psi_u(t) = e^{-j3\pi/8 \cdot t/T}$, respectively, is carried out.
The received signal consists of the sum of U filtered co-channel signals and a white Gaussian noise component. It is assumed that the channels change slowly enough to allow for the assumption that they are constant for the time covered by most of the energy of the transmission impulse. Then, the signal at the receive filter's output is given by

$$\begin{aligned} r(t) &= \sum_{u=1}^{U} V_u\Big(\sum_n c_u(n)\phi_u(n) \int_{-\infty}^{\infty} g_u(\tau - \tau_u, t) \\ &\quad \times x(t - nT - \tau)d\tau + n(t)\Big) \cdot \psi_u(t) \\ &= \sum_{u=1}^{U} \sum_n c_u(n)\phi_u(n)h_u(t - nT, t)\psi_u(t) \\ &\quad + n(t) \cdot \psi_u(t) \end{aligned} \tag{6}$$

with $\phi_u(n)$ the rotation and $\psi_u(t)$ the derotation corresponding to the modulation scheme, and $h_u(\tau, t)$ the composite channel impulse response of the u-th user, given by

$$h_u(\tau, t) = \int_{-\infty}^{\infty} V_u g_u(v - \tau_u, t)x(\tau - v)dv. \tag{7}$$

The relative delay τ_u is viewed as part of the channel impulse response [10]. The derotation in Eq. (6) can also be done after sampling. In this case, a sampled derotation has to be used. $n(t)$ in Eq. (6) is the noise component at the receive filter output. With white noise $z(t)$ we get the auto-correlation function (ACF) of $n(t)$

$$R_n(\alpha) = N_0 v(\alpha) \tag{8}$$

with

$$v(\alpha) = \int_{-\infty}^{\infty} u_R(t) u_R(\alpha - t) dt$$

being the deterministic ACF of the root-Nyquist receive filter. The ACF of $h_u(\tau, t)$ from Eq. (7) gets

$$R_{h_u}(\tau_1, \tau_2, \Delta t) \quad = \quad E\left[h_u(\tau_1, t) h_u^\star(\tau_2, t - \Delta t)\right]. \tag{9}$$

We utilize a wide-sense stationary (WSS) channel. Also, uncorrelated scattering (US) is assumed and a separable scattering function [10]. Thus, Eq. (9) was separable in a delay dependent part $G_u(\tau)$, which can be calculated from the convolution of the power delay profile $P_{g_u}(\tau)$ and the probability density function of the delay τ_u, $f_u(\tau_u)$, and a part depending on the time difference $R_{g_u}(\Delta t)$. If the delay is known, $f_u(\tau_u)$ can be replaced by a Dirac impulse $\delta(\tau - \tau_u)$ [10].

Now, we show the influence of the rotation and derotation on the overall channel impulse response to be estimated. Starting from Eq. (6) and Eq. (7) with $\tau_u = 0$, we get

$$\tilde{h}_u(\tau, t) = \phi_u(\frac{t - \tau}{T}) h_u(\tau, t) \psi_u(t). \tag{10}$$

For EDGE e.g., we get

$$\begin{aligned} \tilde{h}_u(\tau, t) &= e^{j \frac{3\pi}{8} \frac{t-\tau}{T}} h_u(\tau, t) e^{-j \frac{3\pi}{8} \frac{t}{T}} \\ &= e^{j \frac{3\pi}{8} \frac{\tau}{T}} h_u(\tau, t) \end{aligned} \tag{11}$$

It can easily observed in Eq. (11), that the rotation and derotation turns the overall channel impulse response in τ-direction by the factor $e^{j3\pi/8 \cdot (\tau/T)}$.

III. JMSE Channel Estimation

The adaption of the weighting factors requires a suitable error measure. Two measures have established: the mean square error (MSE) and the square error. Minimization of the former leads to the minimum mean square error (MMSE) criterion, of the latter to the least squares (LS) criterion.

If no statistical properties of the transmission channel are known or estimated, the LS approach is taken. The sum of the squared errors is computed over a suitable time interval and is utilized in order to adapt the coefficients

$$\overline{e^2(k)} = \sum_{l=0}^{k} |e(l)|^2. \tag{12}$$

Suitable channel coefficients are chosen so as to minimize this sum. On the other hand, the MSE estimation of the channel coefficients needs the knowledge of expected values, so the statistical properties of the channel such as the first moment and the ACF. Also, the temporal ACF is applied in [15] and [10] to take time variations of the channel into account. The MSE is $E\left[|e(k)|^2\right]$.

A comparison of both approaches can be found in [16].

Here, the conditional mean estimator [16] as utilized in [10] has to be only slightly modified. The interpolation approach has to be skipped for the aforementioned reasons. For this reason, the performance decreases.

Starting from Eq. (6), without derotation, after sampling with $t = kT/2$, we get [10]

$$r(k) = \sum_{u=1}^{U} \underline{c}_u^T(k) \underline{h}_u(k) + n(k), \quad k \in \mathbb{Z}, \tag{13}$$

as the discrete received signal. Here, the data vector contains the rotated data symbols and is given by Eq. (14).

The rotated symbols are given by $\tilde{c}_u(k) = c_u(k)\phi_u(k)$. The u-th user channel coefficient vector is

$$\underline{h}_u(k) = \begin{bmatrix} h_u(0; \frac{kT}{2}) \\ h_u(\frac{T}{2}; \frac{kT}{2}) \\ \vdots \\ h_u(N_h T; \frac{kT}{2}) \\ h_u((N_h + \frac{1}{2})T; \frac{kT}{2}) \end{bmatrix}. \tag{15}$$

$\underline{h}_u(k)$ consists of samples of the causal overall channel impulse response $h_u(\tau, t)$ spaced $T/2$ apart. N_h are the symbols at the beginning of the training sequence being subject to intersymbol interference. If an adaptive filter is used for channel estimation, they can be used for initialization. However, this is not the approach taken here. The length of $\underline{h}_u(k)$ in Eq. (15) is $2N_h$.

The second order statistics of the channel are described by the auto-correlation matrix

$$\underline{R}_{h_u}(\alpha) = E\left[\underline{h}_u(k) \underline{h}_u^H(k - \alpha)\right]. \tag{16}$$

With the ACF from Eq. (9), we get the element in vth column and wth row as

$$\underline{R}_{h_u}(\alpha)_{v,w} = R_{h_u}\left(\frac{vT}{2}, \frac{wT}{2}, \frac{\alpha T}{2}\right), \tag{17}$$

with $v, w \in \{0, \ldots, 2N_h + 1\}$. Although a WSSUS channel is assumed, the elements of $\underline{h}_u(k)$ are correlated, since $\underline{h}_u(k)$ is found from the convolution of $g_u(\tau, t)$ with the CCF of the transmit filter and the receive filter given in Eq. (7).

The impulse responses of the individual users are merged to form a vector of length $2U(N_h + 1)$

$$\underline{h}(k) = \left[\underline{h}_1^T(k)\, \underline{h}_2^T(k) \, \ldots \, \underline{h}_U^T(k)\right]^T \tag{18}$$

to allow for a joint channel estimation. The auto-correlation-matrix of the vector $\underline{h}(k)$ is $\underline{R}_h(\alpha) = E\left[\underline{h}(k)\underline{h}^H(k - \alpha)\right]$. Since the channel vectors are stochastically independent, $\underline{R}_h(\alpha)$ is a block diagonal matrix consisting of $\underline{R}_{h_u}(\alpha)$, $u = 1, ..., U$ on the main diagonal.

The channel estimator uses those samples of the received signal not subject to unknown data symbols. These samples form the vector

$$\underline{r} = \left[r(0)\, r(1) \, \ldots \, r(2(N_s - N_h) - 1)\right]^T. \tag{19}$$

The length of $\underline{r}$ is $2(N_s - N_h)$, with N_s being the length of the training sequence (GSM: $N_s = 26$).

A reduced rank approach through eigenvalue decomposition is utilized [10] in order to reduce the MSE. The auto-correlation

$$\underline{c}_u(k) = \begin{cases} [\; \tilde{c}_u(\lfloor \frac{k}{2} \rfloor) \quad 0 \quad \tilde{c}_u(\lfloor \frac{k}{2} \rfloor - 1) \quad 0 \quad \ldots \quad \tilde{c}_u(\lfloor \frac{k}{2} \rfloor - N_h) \quad 0 \;]^T & \text{, k even} \\ [\; 0 \quad \tilde{c}_u(\lfloor \frac{k}{2} \rfloor) \quad 0 \quad \tilde{c}_u(\lfloor \frac{k}{2} \rfloor - 1) \quad \ldots \quad 0 \quad \tilde{c}_u(\lfloor \frac{k}{2} \rfloor - N_h) \;]^T & \text{, k odd} \end{cases} \tag{14}$$

matrix of the received signal sampled at double symbol rate is given by $\underline{R}_{rr} = E\left[\underline{r}\underline{r}^H\right]$. Here, the QR algorithm is used in order to decompose $\underline{R}_{rr}$. Thus, we get the $P \times P$ matrix $\underline{\Lambda}_1 = diag(\lambda_1, \lambda_2, ..., \lambda_P)$, with $\lambda_i, i = 1, ..., P$ being the P dominant eigenvalues of $\underline{R}_{rr}$. This direct method is used, for it is assumed that the eigenvalue decomposition can be done on the whole matrix on a digital signal processor (DSP). Many state-of-the-art DSPs provide this feature. If not, an iterative method, such as the implicitly restarted Arnoldi method (IRAM), can be utilized. The non-quadratic matrix $\underline{Q}_1$ consists of the eigenvectors $\underline{q}_i, i = 1, ..., P$ belonging to the eigenvalues in $\underline{\Lambda}_1$ arranged in columns. The estimation of the channel vector $\underline{h}(k)$ is done on basis of the low rank model $\underline{w} = \underline{Q}_1^H \underline{r}$ [10] instead of using the vector $\underline{r}$. Note that the components w_i of $\mathbf{w}$ are the coefficients of the truncated Karhunen-Loève expansion:

$$\underline{r} = \sum_{i=1}^{P} w_i \underline{q}_i, \tag{20}$$

minimizing the squared error $\sum_{i=P+1}^{2(N_s-N_h)} E\{|w_i|^2\} \underline{q}_i \underline{q}_i^H$ of the approximation, trading an increased bias against a reduced MSE $E\{||\hat{\underline{h}} - \underline{h}||^2\}$. The reduced vector $\underline{w}$ has the covariance matrix $\underline{R}_w = \underline{Q}_1^H \underline{R}_{rr} \underline{Q}_1 = \underline{\Lambda}_1$.

The optimum MSE estimation of $\underline{h}(k)$ based on $\underline{w}$ has been shown to be given by the conditional mean $\underline{v}(k) = E\left[\underline{h}(k)|\underline{w}\right]$. Since $\underline{h}(k)$ and as $\underline{w}$ are jointly Gaussian, the conditional mean estimate is [10] without interpolation, thus

$$\begin{aligned} \hat{\underline{h}}(k) &= E\left[\underline{h}(k)\underline{w}^H\right] \underline{R}_w^{-1} \underline{w} \\ &= \underline{R}_{hr} \underline{R}_{rr}^{\#} \underline{r}, \end{aligned} \tag{21}$$

with $\underline{R}_{hr}(k) = E\left[\underline{h}(k)\underline{r}^H\right]$ the cross-correlation matrix of the super-vector $\underline{h}(k)$ and the receive vector $\underline{r}$ and $\underline{R}_{rr}^{\#} = \underline{Q}_1 \underline{\Lambda}_1^{-1} \underline{Q}_1^H$.

IV. Interference Scenarios

In [17] and [18], different scenarios are suggested in order to assess the implications of interference on the system performance. Two important measures in conjunction with IC are the Carrier-to-Interference Ratio (CIR)

$$CIR = 10 \log_{10} \left(\frac{C}{\sum_k I_k} \right) \tag{22}$$

and the Dominant-to-Rest-of-Interference Ratio (DIR)

$$DIR = 10 \log_{10} \left(\frac{I_{max}}{\sum_k I_k - I_{max}} \right), \tag{23}$$

with C the power of the carrier signal, I_k the power of the k-th interferer and I_{max} the power of the strongest interferer. Eq. (22) describes the ratio between the power of the desired signal and the total signal power of the interferers, Eq. (23) describes the ratio of the strongest co-channel interferer's power to the power of the remaining interferers. In [19] a maximum of two interferers is suggested.

A dominant co-channel interferer typically occurs in GSM networks with tight frequency reuse. If the fractional load increases, the probability of a second interferer increases. Adjacent channel interference, which is also supposed to be addressed by ARP, is not considered in this paper.

In addition to a pure GMSK and 8-PSK scenario, desired and co-channel users occur with GMSK and 8-PSK modulation schemes at the same time and have to be considered in a mixed environment [19].

Also, synchronized as well as unsynchronized networks are considered. Delays between co-channel signal components in the received signal are due to different delays on the channel. According to [19], with reuse 1/1 and a cell width of 3km, delays of less than 3 symbols occur with 60% probability, and a maximum delay of 3 symbols is realistic. If the base stations are not synchronized, up to two bursts per interferer overlap with the desired user's burst. In this paper, we assumed that either up to two interfering training sequences are overlapping with the desired user's training sequence. This condition is very weak, it is more than covered by the assumption of a maximum delay of 5.5 symbols. Both consecutive bursts of the interferer are supposed to have equal power.

V. Numerical Results

In [10], instead of the estimator in Eq. (21), an interpolation over multiple bursts is suggested. This is only possible in GSM/EDGE networks when operating in GPRS/EGPRS mode and frequency hopping is switched off, since otherwise there are no neighboring time slots assigned to a user, and the mobile channel gets close to the coherence time even at moderate velocities, if the slots taken into consideration are one frame apart. Also, if frequency hopping is used, the mobile channel is completely different for a user's channel to be estimated, from frame to frame. Therefore, no interpolation can be utilized, degrading the performance in terms of MSE.

With $\underline{c}(i) = [\underline{c}_1(i)\underline{c}_2(i) \ldots \underline{c}_U(i)]^T$ containing only the training information of the U users, the j-th column of $\underline{R}_{hr}(k)$ can be easily shown to be $\{\underline{R}_{hr}(k)\}\,|_j = \underline{R}_h(k-j)\underline{c}(j)^*$ and the i-th row, j-th column entry of $\underline{R}_{rr}$ gets $\{\underline{R}_{rr}\}\,|_{i,j} = \underline{c}(i)^T \underline{R}_h(i-j)\underline{c}(i)^* + R_n(\alpha)\,|_{\alpha=\frac{iT}{2}}$ for the simplified version of [10].

In order to calculate the statistics, the channel estimator needs to know $\underline{c}(i)$, containing the training sequences of the interfering users. If they are different from the desired user in a downlink scenario, they might be estimated from the seven possible remaining sequences. This topic, as well as their relative position to the desired burst is treated in [9].

By partial subtraction of the desired user's channel from the received signal, the rate of detection is enhanced dramatically. Also, the modulation has to be estimated.

The channel estimator needs to know the ACF of the channel and the ACF of the noise. Assuming a given scattering environment, the ACF depends on the Doppler frequency and the power delay profile. In [15], it is shown that an estimator of this type is robust to changes in the power delay profile, when the delay spread is assumed higher or equal to the actual delay spread. Here, the delay dependent part of $G_u(\tau)$ and $\underline{R}_{\underline{h}_u}$ are calculated on the basis of a typical urban (TU) channel profile. $\underline{R}_{rr}^{\#}$ is calculated as $\underline{R}_{rr}^{\#} = \sum_{i=1}^{P} \frac{1}{\lambda_i} \underline{q}_i \underline{q}_i^H$ after applying QR algorithm to $\underline{R}_{rr}$. For evaluation of the efficiency of the Joint MSE Channel Estimation several simulations are made in the downlink. The simulations contain a comparison between interference signals with different power and modulation types. The efficiency of the channel estimation is assessed by the estimation error $\underline{e}(k) = \underline{h}(k) - \hat{\underline{h}}(k)$ of the MSE estimator. The estimation error covariance matrix is derived in [10] to be

$$\underline{R}_e(k) = \underline{R}_h(0) - \underline{R}_{hr}(k)\underline{R}_{rr}^{\#}\underline{R}_{hr}^H(k) \tag{24}$$

and the MSE is the u-th block along the main diagonal of $\underline{R}_e(k)$, divided by the energy of the u-th channel:

$$\sigma_{e_u}^2(k) = \frac{trace(\underline{R}_{e_u}(k))}{trace(\underline{R}_{h_u}(0))}. \tag{25}$$

The ideal channel estimation would result in $\sigma_{e_u}^2(k) = 0$.

The carrier frequency is set to $900\ MHz$ and the velocity of the mobile station was $50\ km/h$. This leads to a maximal Doppler frequency of $41.67Hz$ and a coherence time of $12ms$. Ideal frequency hopping is assumed.

The root-Nyquist receive filter has a roll-off of 0.3. This is an academic choice. In realized systems, the receive filter may be optimized to reject adjacent channel interference and is in general dependent on the AD-converter, its anti-aliasing low-pass filter and the conversion technique.

A. Dependency on E_b/N_0, CIR and DIR

Fig. 4(a) shows the dependency of $\sigma_{e_u}^2(k)$ on the CIR. The higher the difference of the interferers in power, the better the estimate of the channel. Fig. 4(b) shows $\sigma_{e_u}^2(k)$ for the interferer in the same environment, depending on the CIR.

From Figs. 5(a) and 5(b) it can be seen that a better estimate of the channel results if a dominant interferer exists. The best results can be achieved with a single interferer, i.e. $DIR = \infty$. For the estimation of the channel impulse responses of the interfering signals the error for both interfering signals is equal if the transmission power is equal ($DIR = 0\ dB$). If there is a dominant interferer its estimation becomes better, whereas the estimation for the weak signal degrades seriously.

B. Dependency on the Relative Delays

As already said, delays occurring due to different distances to the emitting base stations are covered in the estimator. Fig. 3 shows a scenario when the interfering signal arrives at the mobile station after the desired one, which is the case of highest probability due to the geometry. The delay is normalized to a symbol interval. It can be seen that $\sigma_{e_u}^2(k)$ increases if the delay between the desired signal and the interferer gets higher than 2.5 times the symbol interval. Furthermore the part of the channel impulse response caught by the matrix in Eq. (15) becomes smaller.

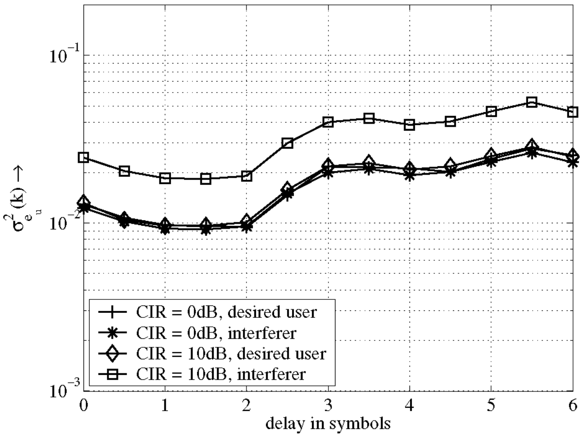


Fig. 3. $\sigma_{e_u}^2(k)$ depending on relative time shift, DIR=∞, $E_b/N_0 = 20dB$.

VI. Conclusion

It is observable that $\sigma_{e_u}^2(k)$ of the desired user is insensitive to changes in DIR. If the JMSE is to be utilized with a JDDFSE, the high $\sigma_{e_u}^2(k)$ for the interferer at higher CIR might seriously degrade the BER performance. Differences between GMSK and 8-PSK could not be observed, which is due to their temporal ACF, which is roughly the same. When comparing the principal results with [9], the JMSE technique seems to be more sensitive to relative time shifts.

VII. Acknowledgments

The authors would like to thank Prof. James K. Cavers of Simon Fraser University, Barnaby, Canada, and Stephen J. Grant, Ericsson Research, North Carolina, U.S.A., for their helpful comments on the topic.

References

[1] J. H. Winters, "Optimum combining in digital mobile radio with cochannel interference," *IEEE J. Select. Areas Commun., vol. SAC-2., pp. 528-539, July 1984.*

[2] K. Giridhar, "Nonlinear techniques for the joint estimation of cochannel signals," *IEEE Trans. Commun., vol. 45 no. 10, pp. 473-484, April 1997.*

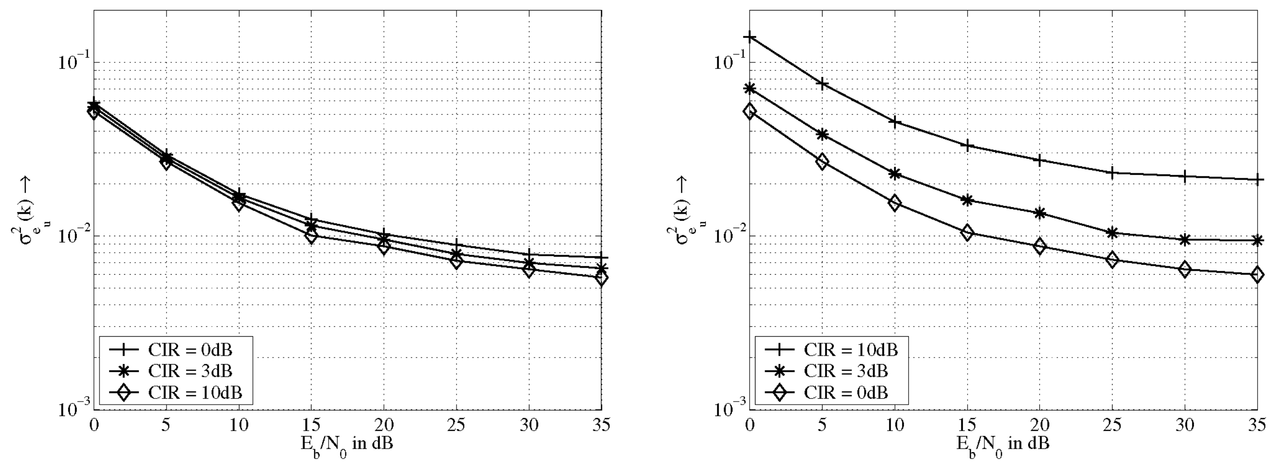
[3] Joung, J. and Stüber, G. L., "Performance of truncated co-channel interference canceling MLSE for TDMA systems," *Proc. IEEE Veh. Technol. Conf. '98, Ottawa, Canada, May 18-21, 1998, pp. 1710-1714.*

[4] Kim, K. and Stüber, G. L., "Interference cancelling receiver for range extended reception in tdma cellular systems," *Proc. IEEE Veh. Technol. Conf. (VTC), Fall 2003.*

[5] Hui, D. and Ramesh, R., "Maximum likelihood sequence estimation in the presence of constant envelope interferers," *Proc. IEEE Veh. Technol. Conf. (VTC), Fall 2003.*

[6] 3GPP TSG GERAN, *TDOC 031964. Available at: http://www.3gpp.org.*

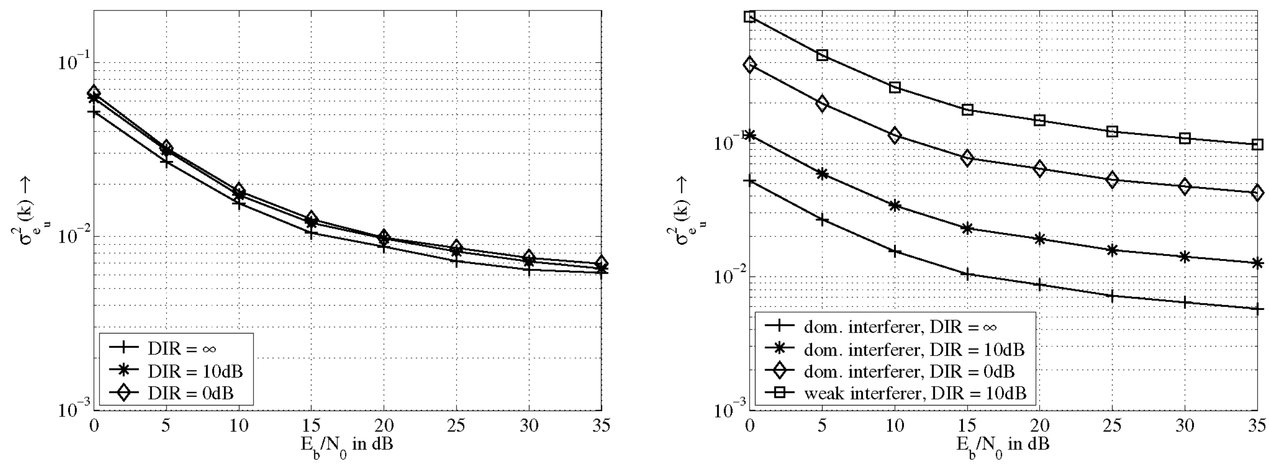
[7] Glisic, S. G. and Pirinen, P., *Cochannel Interference in digital cellular TDMA Networks, in: Wiley Encyclopedia of Telecommunications, pp. 448-466,* John Wiley & Sons, New York, 1. edition, 2003.



(a) $\sigma^2_{e_u}(k)$ for desired user depending on CIR, DIR=∞.

(b) $\sigma^2_{e_u}(k)$ for interfering user depending on CIR, DIR=∞.

Fig. 4. Estimation error $\sigma^2_{e_u}(k)$.



(a) $\sigma^2_{e_u}(k)$ for desired user depending on DIR, CIR=$0dB$.

(b) $\sigma^2_{e_u}(k)$ for interfering user depending on DIR, CIR=$0dB$.

Fig. 5. Estimation error $\sigma^2_{e_u}(k)$.

[8] B. Hagerman, "Downlink relative co-channel interference powers in cellular radio systems," *Proc. IEEE Veh. Technol. Conf.*, 1995.

[9] F. Nordström, *Interferer Detection and Cancellation in TDMA-Systems*, Licentiate thesis, LUTFMS-2004-2001, Centre for Mathematical Sciences, Lund University, Sweden, Nov, 2001.

[10] Grant, S. J. and Cavers, J. K., "Multiuser channel estimation for detection of cochannel signals," *IEEE Trans. Commun., vol. 49 no. 10, pp. 1845-1855, October 2001.*

[11] P. A. Laurent, "Exact and approximate construction of digital phase modulations by superposition of amplitude modulated pulses (AMP)," *IEEE Trans. Commun., vol. COM-34 no. 2, pp. 150-160, February 1986.*

[12] A. Baier, "Derotation techniques in receivers for MSK-type CPM signals," *Proc. Fifth. Europ. Sig. Proc. Conf. (EUSIPCO-90), Barcelona pp. 1799-1802, Sept. 1990.*

[13] Murphy, P. A. and Ford, G. E., "Cochannel receivers for CPM signals based upon the laurent representation," *Proc. Sixth Symposium on Wireless Personal Communications, Blacksburg, VA, June 5-7, 1996.*

[14] Gerstacker, W. and Schober, R., "Equalization concepts for EDGE," *IEEE Trans. Wireless Commun., Vol. 1, No. 1, pp. 190 -199, January 2002.*

[15] J. K. Cavers, "An analysis of pilot assisted modulation for Rayleigh fading channels," *IEEE Trans. Veh. Technol., 40(4), pp. 686-693, Nov. 1991.*

[16] S. Haykin, *Adaptive Filter Theory*, Prentice Hall, Englewood Cliffs, New Jersey, 4. edition, 1983.

[17] 3GPP TSG GERAN, *Tdoc GAHS-030002, Single Antenna Interference Cancellation – Evaluation Principles and Scenarios, SAIC ad hoc #1, Atlanta, Source: Ericsson, Jan. 2003.*

[18] 3GPP TSG GERAN, *Tdoc GAHS-030005, Scenarios and Modelling Assumptions for SAIC in GERAN, SAIC ad hoc #1, Atlanta, Source: Siemens, Jan. 2003.*

[19] 3GPP TSG RAN, *3G TS 25.xxx. Available at: http://www.3gpp.org.*