CONSISTENT ESTIMATION OF RAYLEIGH FADING CHANNEL SECOND ORDER STATISTICS IN THE CONTEXT OF THE WIDEBAND CDMA MODE OF THE UMTS.

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ABSTRACT

In this paper, we address the problem of second order statistics estimation of a selective Rayleigh channel in the context of the wideband CDMA mode of the UMTS. The data to be transmitted are sent over slots over which the channel is assumed to remain constant. Each slot contains a pilot symbol sequence from which the channel can be estimated. The covariance matrix of the channel is usually estimated by a denoised version of the empirical covariance matrix of the trained channel estimate. However, this estimate is not consistent in the UMTS context. In this paper, we propose a new consistent estimate of the channel covariance matrix, and evaluate the performances of two Wiener like channel estimation schemes based on the proposed estimate.

1. INTRODUCTION

In the context of high rate mobile communication systems, the received signal is often corrupted by a fading frequency selective channel. In this case, the coefficients of the equivalent discrete-time channel can be considered as highly low pass time-varying centered Gaussian random variables (see e.g. [7]) which must be estimated in order to retrieve the transmitted data. In practice, the data to be transmitted are sent over slots on which the channel coefficients can be considered as constant, and containing a training sequence from which the channel coefficients are estimated by a least-squares or a correlation procedure ([1]). The accuracy of these estimates, which depends of course on both the length of the training sequence and on the signal to interference plus noise ratio, may have an important influence on the global performance of the receiver. This turns out to be the case in the context of the wideband CDMA mode of the third mobile generation system (UMTS). In the downlink, the size of the training sequence is rather short, and the accuracy of the conventional channel estimate is very poor when the system is heavily loaded. This affects significantly the performances of most of the conventional receivers based on this channel estimate.

In order to improve the performances of the channel estimate, one can use semi-blind approaches. These methods aim at estimating the channel not only from the observations corresponding to the transmission of the training sequence, but also from the entire slot. However, the existing algorithms have a very high computational cost, especially in the context of multi-users systems (see e.g. [4],[3]). This paper is devoted to a completely different approach. We denote by $\mathbf{h}_m$ the vector of the coefficients of the discrete time equivalent channel on slot number m, and assume that each vector $\mathbf{h}_m$ can be interpreted as a zero-mean slot-varying, i.e. $\mathbf{h}_m \neq \mathbf{h}_{m'}$ if $m \neq m'$, Gaussian random vector. However, the probability distributions of the vectors $(\mathbf{h}_m)_{m=0}^{M-1}$ can be considered as slot-invariant, at least if the number of considered slots M is not too large enough. If the covariance matrix $\mathbf{\Gamma}$ of this distribution were known, it could be possible to use a Wiener estimate of $\mathbf{h}_m$. In order to explain this, assume for the moment that the conventional trained estimate $\hat{\mathbf{h}}_m$ can be written as

$$\hat{\mathbf{h}}_m = \mathbf{h}_m + \boldsymbol{\epsilon}_m \quad (1)$$

where $\boldsymbol{\epsilon}_m$ is a random vector independent of $\mathbf{h}_m$ with known covariance matrix $\mathbf{\Sigma}$. In this case, the classical Wiener estimate, given by $E(\mathbf{h}_m\hat{\mathbf{h}}_m^H)\left(E(\hat{\mathbf{h}}_m\hat{\mathbf{h}}_m^H)\right)^{-1}\hat{\mathbf{h}}_m = \mathbf{\Gamma}(\mathbf{\Gamma}+\mathbf{\Sigma})^{-1}\hat{\mathbf{h}}_m$ may produce significant improvement. Wiener channel estimation seems to have been introduced by [2] in the context of mono-user system (the GSM system). However, we note that it can be considered as a simplification of Kalman procedures developed in the context of fast fading channel estimation ([10]), in which the channel coefficients cannot be assumed to be constant over the duration of a slot. In the context of GSM system considered in [2], relation (1) is satisfied, and the covariance matrix $\mathbf{\Sigma}$ of $\boldsymbol{\epsilon}_m$ can be assumed to be a multiple $\sigma^2 I$ of the identity matrix. As $\mathbf{\Gamma}$ is of course unknown, [2] proposed to estimate it by

$$\hat{\mathbf{\Gamma}} = \frac{1}{M}\sum_{m=0}^{M-1}\hat{\mathbf{h}}_m\hat{\mathbf{h}}_m^H - \sigma^2 I \quad (2)$$

if σ^2 is known. If σ^2 is unknown, these authors propose to estimate it by the smallest eigenvalue of matrix $\frac{1}{M}\sum_{m=0}^{M-1}\hat{\mathbf{h}}_m\hat{\mathbf{h}}_m^H$. (2) is consistent as soon as $\mathbf{\Gamma}$ is rank deficient, a condition which is often met in practice when the channel is supposed to be specular. The estimate (2) of $\mathbf{\Gamma}$ turns out to be consistent (in the sense that $\hat{\mathbf{\Gamma}}$ converges toward $\mathbf{\Gamma}$ when $M \to \infty$) if relation (1) holds, and if the covariance matrix of $\boldsymbol{\epsilon}_m$ is a multiple of the identity matrix. These conditions are however not verified in the context of the wideband

CDMA mode of the UMTS (see below). The purpose of this paper is twofold. We first propose a new consistent estimation scheme of the matrix $\mathbf{\Gamma}$ in the context of the downlink of the wideband CDMA mode of the UMTS. Next, we study and compare the performances of two channel estimation algorithms (Wiener and rank reduction Wiener) using our consistent estimation scheme of $\mathbf{\Gamma}$.

This paper is organized as follows. In section 2, we precise the structure of the signals that are transmitted and received in the downlink of the wideband CDMA mode of the UMTS. In section 3, we present our consistent estimate of $\mathbf{\Gamma}$. We study the corresponding estimation schemes in section 4, and evaluate their performances by numerical simulations in section 5.

2. THE DOWNLINK UMTS SIGNAL STRUCTURE

2.1. The UMTS specifications

We consider a mobile station which is supposed to receive slots of QPSK data symbols sequence $(b_{m,0}(l))_{l=0,K}$ transmitted by the base station of its closest cell. Here, the subscript m represents the index of the slot, l represents the index of the symbol of the slot m, K is the number of symbols per slot. The base station transmits simultaneously Q other data symbol slots $(b_{m,q})_{q=1,\dots Q}$ to Q other users.

We first precise the structure of the signal received by mobile station 0. In the context of UMTS, different users may use different spreading factors. In order to simplify the notations of this paper, we assume that the same spreading factor N is assigned to the $(Q+1)$ users of the cell under consideration. The number of chips per slot is thus equal to NK. The reader may check that this assumption does not induce any restriction and that our results remain valid if different spreading factors are assigned to certain users. Each sequence of symbols $b_{m,q}(l)$ is spread by a BPSK periodic sequence of period N $c_q(n)$. The corresponding $(Q+1)$ chip sequences are finally scrambled by the same long aperiodic code (this code characterizes the cell). We denote by $s_m(n)$ the value of the scrambling code on chip n of slot m. In the following, we denote by $d_{m,q}(n)$ the chip sequence corresponding to slot m of user q, which according to the above specifications, is given by $d_{m,q}(lN+k) = b_{m,q}(l)c_q(k)s_m(lN+k)$ for $0 \leq k \leq N-1$ and $0 \leq l \leq K-1$.

The continuous-time signal $x_a(t)$ received by mobile station 0 and corresponding to the transmission of slot m of the various users is thus given by $x_{m,a}(t) = \sum_{q=0}^{Q} \mu_q \sum_n d_{m,q}(n) h_{m,a}(t - nT_c) + w_a(t)$. Here, T_c represents the chip period, $h_{m,a}(t)$ represents the (unknown) impulse response resulting from the shaping filter (i.e. a square root raised cosine of roll-off 0.22), the propagation channel between the base station and the mobile station 0, and the reception filter. We assume without restriction that it is causal. Note that it depends on the slot m to take into account the time variations of the propagation channel. The coefficients $\mu_0, \dots, \mu_Q$ are positive, and represent the square roots of the powers of the contributions of each active user to the received signal. In the following, we assume without loss of generality that $\mu_0 = 1$. The coefficients $(\mu_q)_{q=1,\dots,Q}$ thus represent the relative powers of the other users. Finally, $w_a(t)$ is an additive noise due to the signals emitted by other interfering cells and to the background noise assumed to be white Gaussian with spectral density $N_0/2$. We assume that the mobile station 0 has synchronized with the base station. This implies in particular that the mobile has a perfect knowledge of the scrambling code sequence. On the other hand, each slot contains a pilot sequence of P symbols which can be used in order to estimate the channel. In other words, the mobile station 0 knows the first NP chips of each slot m transmitted by user 0 (i.e. the sequence $d_{m,0}(0), \dots, d_{m,0}(NP-1)$). However, the mobile station 0 is not aware of the pilot sequences transmitted by the users $1, \dots, Q$.

2.2. The discrete-time equivalent model

The signal $x_a(t)$ is sampled at the period $T_c/2$. We denote by $\mathbf{x}_m(n)$ the two-dimensional vector $\mathbf{x}_m(n) = (x_{m,a}(nT_c), x_{m,a}(nT_c + T_c/2))^T$ and by $\mathbf{h}_m(k)$ the vector $\mathbf{h}_m(k) = (h_{m,a}(kT_c), h_{m,a}(kT_c + T_c/2))^T$. We put $\mathbf{h}_m = (\mathbf{h}_m(0)^T, \dots, \mathbf{h}_m(L)^T)^T$ where LT_c represent the maximum duration of the channel. It is easily seen that the discrete-time signal $\mathbf{x}_m(n)$ can be written as $\mathbf{x}_m(n) = \sum_{k=0}^{L} \sum_{q=0}^{Q} \mu_q \mathbf{h}_m(k) d_{m,q}(n-k) + \mathbf{w}_m(n)$ where $\mathbf{w}_m(n)$ is defined as $\mathbf{x}_m(n)$. We now formulate the following assumptions :

- **(A1)** For each m, $\mathbf{h}_m$ is a complex Gaussian random vector, and its covariance matrix is time invariant, i.e. it does not depend on m. In the following, we denote by $\mathbf{\Gamma} = E(\mathbf{h}_m \mathbf{h}_m^H)$ this covariance matrix.
- **(A2)** The (known) sequence $(s_m(n))_{n=0,\dots,NK-1,\ m\in\mathbb{Z}}$ is assumed to coincide with a realization of an independent identically distributed centered QPSK sequence. In particular, for each p and each function Φ, $\lim_{M\to+\infty} \frac{1}{M} \sum_{m=0}^{M-1} \Phi(s_m(n+\tau_1), \dots, s_m(n+\tau_p)) = E_s(\Phi(s(\tau_1), \dots, s(\tau_p)))$ where s represents a centered QPSK i.i.d. sequence.
- **(A3)** For $q \geq 0$, the symbol sequence transmitted by user q is i.i.d. The various sequences are also mutually independent.

3. ESTIMATION OF THE CHANNEL COVARIANCE MATRIX

The conventional estimate is obtained by correlating the received signal with delayed versions of the chip sequence corresponding to the pilot symbols sequence : $\hat{\mathbf{h}}_m(k) = \frac{1}{NP} \sum_{n=0}^{NP-1} \mathbf{x}_m(n+k) d_{m,0}^*(n)$. Using the expression of $\mathbf{x}_m(n+k)$, we get that the estimation error $\boldsymbol{\epsilon}_m = \hat{\mathbf{h}}_m - \mathbf{h}_m$ has three components. The first one is the contribution of the auto-correlations of $d_{m,0}$ given by $\frac{1}{NP} \sum_{l=0}^{L} \mathbf{h}_m(l) \sum_{n=0}^{NP-1} d_{m,0}(n+k-l) d_{m,0}^*(n) - \mathbf{h}_m(k)$, the second is the interference of other users of the cell: $\frac{1}{NP} \sum_{q=1}^{Q} \mu_q \sum_{l=0}^{L} \mathbf{h}_m(l) \sum_{n=0}^{NP-1} d_{m,q}(n+k-l) d_{m,0}^*(n)$ and the third is $\frac{1}{NP} \sum_{n=0}^{NP-1} \mathbf{w}_m(n+k) d_{m,0}^*(n)$. Our problem thus differs deeply from the context used by [2], where the first component is zero and the second one does not exist (see (1), and the corresponding assumptions). Moreover, we observe that vectors $\mathbf{h}_m$ and $\boldsymbol{\epsilon}_m$ are not statistically independent and that the covariance matrix of $\boldsymbol{\epsilon}_m$ is not a multiple of the identity matrix: it actually depends on $\mathbf{\Gamma}$, and on the unknown distribution of the co-cell interference. Moreover, due to the scrambling code, it is not time-invariant, i.e. it depends on the slot under consideration. This shows that the standard estimate (2) is not consistent in the present context, and that its performance may be very poor if the multi-user interference and the co-cell interference terms are dominant. A quite different approach is thus needed to estimate matrix $\mathbf{\Gamma}$.

In order to present the core of our new estimation method, we need to introduce some notations. First, we denote by

$\mathbf{\Delta}_m$ the covariance matrix $E(\hat{\mathbf{h}}_m \hat{\mathbf{h}}_m^H)$ of vector $\hat{\mathbf{h}}_m$, which as shown below, depends on m. Next, we denote $\mathbf{\Delta}_\infty$ the "temporal mean" of matrices $\mathbf{\Delta}_m$ defined by $\mathbf{\Delta}_\infty = \lim_{M\to\infty} \frac{1}{M}\sum_{m=0}^{M-1} \mathbf{\Delta}_m$. From now, we denote by $\mathbf{x}_{m,L}(n)$ the $2(L+1)$–dimensional vector $\mathbf{x}_{m,L}(n) = (\mathbf{x}_m(n)^T, \dots, \mathbf{x}_m(n+L)^T)^T$ and put $\mathbf{R}_{x,m} = E(\mathbf{x}_{m,L}(n)\mathbf{x}_{m,L}(n)^H)$ and $\mathbf{R}_{x,\infty} = \lim_{M\to\infty} \frac{1}{M}\sum_{m=0}^{M-1} \mathbf{R}_{x,m}$. Vector $\mathbf{w}_{m,L}(n)$ and matrices $\mathbf{R}_{w,m}$ and $\mathbf{R}_{w,\infty}$ are defined similarly. Our approach is based on the following identities :

Proposition 1 *The matrices $\mathbf{R}_{x,\infty}$ and $\mathbf{\Delta}_\infty$ can be written :*

$$\mathbf{R}_{x,\infty} = (\sum_{q=0}^{Q} \mu_q^2)\mathcal{T}(\mathbf{\Gamma}) + \mathbf{R}_{w,\infty} \tag{3}$$

and

$$\mathbf{\Delta}_\infty = \mathbf{\Gamma} + \frac{1}{NP}(\sum_{q=0}^{Q} \mu_q^2)\,(\mathcal{T}(\mathbf{\Gamma}) - \mathbf{\Gamma}) + \frac{1}{NP}\mathbf{R}_{w,\infty} \tag{4}$$

where matrix $\mathcal{T}(\mathbf{\Gamma})$ represents the block Toeplitz matrix whose each 2×2 block $\mathcal{T}(\mathbf{\Gamma})(k,l)$ is given by $\mathcal{T}(\mathbf{\Gamma})(k,l) = \sum_{(i,j), i-j=k-l} \mathbf{\Gamma}(i,j)$.

It turns out that

$$\mathbf{\Delta}_\infty - \frac{1}{NP}\mathbf{R}_{x,\infty} = (1 - \frac{1}{NP}(\sum_{q=0}^{Q} \mu_q^2))\mathbf{\Gamma} \tag{5}$$

Under certain standard mixing assumptions on sequence $(\mathbf{h}_m)_{m\in\mathbb{Z}}$, matrices $\mathbf{\Delta}_\infty$ and $\mathbf{R}_{x,\infty}$ can be consistently estimated by $\hat{\mathbf{\Delta}}_\infty = \frac{1}{M}\sum_{m=0}^{M-1} \hat{\mathbf{h}}_m \hat{\mathbf{h}}_m^H$ and by $\hat{\mathbf{R}}_{x,\infty} = \frac{1}{MNK}\sum_{m=0}^{M-1}\sum_{n=0}^{NK-1} \mathbf{x}_{m,L}(n)\mathbf{x}_{m,L}(n)^H$ Therefore, relation (5) provides a direct way to estimate $\mathbf{\Gamma}$ consistently up to a constant multiplicative factor by the matrix $\hat{\mathbf{\Gamma}}$ defined by :

$$\hat{\mathbf{\Gamma}} = \hat{\mathbf{\Delta}}_\infty - \frac{1}{NP}\hat{\mathbf{R}}_{x,\infty} \tag{6}$$

Note that proposition 1 can be interpreted as a generalization to random time-varying channels and to the context of the UMTS of the results of [9], where it is shown that a time invariant channel can be deduced from the difference between the covariance matrix build from the observation before and after despreading.

4. IMPROVEMENT OF CHANNEL ESTIMATION USING THE CHANNEL COVARIANCE MATRIX

4.1. Modified Wiener estimation

The classical Wiener channel estimator of $\mathbf{h}_m$ is defined as the orthogonal projection of $\mathbf{h}_m$ over the space generated by the components of the observed random vector $\hat{\mathbf{h}}_m$. It is thus given by

$$\mathbf{h}_m/\hat{\mathbf{h}}_m = E(\mathbf{h}_m\hat{\mathbf{h}}_m^H)(E(\hat{\mathbf{h}}_m\hat{\mathbf{h}}_m^H))^{-1}\hat{\mathbf{h}}_m = E(\mathbf{h}_m\hat{\mathbf{h}}_m^H)\mathbf{\Delta}_m^{-1}\hat{\mathbf{h}}_m$$

where $/$ stands for the usual orthogonal projection operator in the space of finite second order moments random variables. However, this channel estimator cannot be implemented in practice because it is impossible to estimate consistently matrix $\mathbf{\Delta}_m$. We therefore propose to use a modified Wiener estimate defined by $\mathbf{A}_{opt}\hat{\mathbf{h}}_m$, where matrix $\mathbf{A}_{opt}$ minimizes the cost function $\lim_{M\to\infty}\frac{1}{M}\sum_{m=0}^{M-1} E(\|\mathbf{h}_m - \mathbf{A}\hat{\mathbf{h}}_m\|^2)$ The optimal matrix $\mathbf{A}_{opt}$ can be shown to coincide with $\mathbf{A}_{opt} = \mathbf{\Gamma}\mathbf{\Delta}_\infty^{-1}$ and can be consistently estimated by matrix $\hat{\mathbf{\Gamma}}\hat{\mathbf{\Delta}}_\infty^{-1}$. Our modified Wiener channel estimate is thus the vector $\overline{\mathbf{h}}_m$ given by $\overline{\mathbf{h}}_m = \hat{\mathbf{\Gamma}}\hat{\mathbf{\Delta}}_\infty^{-1}\hat{\mathbf{h}}_m$

4.2. Rank reduction of channel subspace

In practice, the performance of the estimate $\overline{\mathbf{h}}_m = \hat{\mathbf{\Gamma}}\hat{\mathbf{\Delta}}_\infty^{-1}\hat{\mathbf{h}}_m$ may be very far from those of the true modified Wiener estimate $\mathbf{\Gamma}\mathbf{\Delta}_\infty^{-1}\hat{\mathbf{h}}_m$. This is in particular the case when the number of slots M used to estimate matrices $\mathbf{\Gamma}$ and $\mathbf{\Delta}_\infty$ is not large enough compared to the dimension of the matrices to be estimated: in this case, the estimates $\hat{\mathbf{\Gamma}}$ and $\hat{\mathbf{\Delta}}_\infty$ are not accurate enough. Fortunately, the performance of the estimate $\overline{\mathbf{h}}_m$ can be improved significantly if matrix $\mathbf{\Gamma}$ is rank deficient (or close to be rank deficient) which turns out to be the case in the context of the so-called multi-path Clarke model ([5]). Let then denote by r its rank. In this case, the channel $\mathbf{h}_m$ can be written as $\mathbf{h}_m = \mathbf{U}\mathbf{g}_m$ where $\mathbf{U}$ represents the matrix build from the r eigenvectors associated to the non zero eigenvalues of $\mathbf{\Gamma}$, and where $\mathbf{g}_m$ is a r–dimensional vector.

Let us first assume that $\mathbf{U}$ is known. The estimation of vector $\mathbf{h}_m$ reduces to the estimation of the r components of $\mathbf{g}_m$, which is an easier problem if r is significantly smaller than the number of components $2(L+1)$ of $\mathbf{h}_m$. $\mathbf{g}_m$ can be estimated by means of a modified Wiener estimate $\tilde{\mathbf{g}}_m$ based on $\hat{\mathbf{g}}_m = \mathbf{U}^H\hat{\mathbf{h}}_m$. However, matrix $\mathbf{U}$ is of course unknown, and replaced in practice in the above procedure by the matrix $\hat{\mathbf{U}}$ of the eigenvectors associated to the r greatest eigenvalues of matrix $\hat{\mathbf{\Gamma}}$.

5. SIMULATIONS

In order to simulate the propagation channel, we have used a realistic simulator ([6]) developed by the research center of France Telecom. We have chosen a 3 paths channel with time-varying complex amplitude corresponding to a mobile speed of 5 Km/h. The spreading factor of the user of interest (i.e. the user 0) is 256. Each slot thus contains 6 useful QPSK symbols, and 4 QPSK pilot symbols (see [1]) which are assumed to be sent with the same power. We take $Q = 55$ other users, so that the load of the cell is $\rho \simeq \frac{1}{5}$. The co-cell interference is absent. In order to compare the statistical performance of the various estimators of the channel, we evaluate the bit error rate corresponding to a conventional RAKE receiver based on the channel estimates. In other words, the decision on symbol $b_{m,0}(l)$ is based on the argument of $\hat{b}_{m,0}(l) = \frac{1}{N}\sum_{k=0}^{L}\sum_{n=0}^{N-1} \mathbf{f}_m^H \mathbf{x}_m(lN+n+k)c_0(n)s_m(lN+n)^*$ where $\mathbf{f}_m$ represents one of the possible channel estimates.

In figure 1, we compare the conventional pilot based estimate $\hat{\mathbf{h}}_m$, the true modified Wiener channel estimate $\mathbf{\Delta}_\infty^{-1}\mathbf{\Gamma}\hat{\mathbf{h}}_m$, the estimated modified Wiener channel estimate based on the estimate (2) of Γ, and the estimated modified Wiener estimate based on the proposed estimation procedure of Γ (6). The performance of the Rake receiver associated with the true channel is also represented. Here, the number of slots used to estimate the various matrices is $M = 240$, which, in the context of the UMTS, corresponds to a duration of 160 ms. We take $L = 20$ chips for the channel size.

We observe that the performance of the true modified Wiener channel estimate is close to that of the true channel. If $E_b/N_0 > 8dB$, the proposed estimate significantly outperforms the approach of (2) because (2) is not a consistent estimate of $\mathbf{\Gamma}$. Note in particular that if $E_b/N_0 > 12dB$, (2) behaves like the conventional trained estimate. Nevertheless, both estimation procedures of $\mathbf{\Gamma}$ are far from the true modified Wiener channel estimate. Fortunately, rank reduction procedures allow to improve a lot the performance. This claim is illustrated in figure 2 where we compare the performance of the proposed reduced rank estimate (6) of $\mathbf{\Gamma}$ and the reduced rank estimate based on the eigendecomposition of (2). In both cases, the rank of the estimate of $\mathbf{\Gamma}$ is evaluated by the procedure proposed in [8], i.e. $\hat{r} = argmin_i \frac{|\hat{\lambda}_i|}{|\hat{\lambda}_{i+1}|}$ where $\hat{\lambda}_1 \geq \hat{\lambda}_2 \geq \cdots \geq \hat{\lambda}_{2L+2}$ are the eigenvalues of the estimate of $\mathbf{\Gamma}$ arranged in decreasing order.

Again, our proposal outperforms a lot the standard procedure based on (2). In particular, if $E_b/N_0 > 16dB$, the performance of the proposed rank reduced estimate is closed to that of the true modified Wiener estimate. However, if $E_b/N_0 < 12dB$, this is no longer the case. Apparently, this is because the detected value of r is 1 in most cases (instead of 3) for those SNR, thus inducing a loss of performance. In order to overcome this drawback, we also plot the performance corresponding to the rank determination procedure consisting in estimating r by $\bar{r}$ defined as the number of positive eigenvalues of (6). We observe an important improvement. Note also that this last procedure is not applicable to the conventional estimate (2).

6. CONCLUSION

In this paper, we have addressed the problem of estimating consistently the covariance matrix $\mathbf{\Gamma}$ of the discrete-time version of a Rayleigh fading channel in the context of the WCDMA mode of UMTS. Our estimate is based on the observation that $\mathbf{\Gamma}$ can be obtained by substracting the temporal mean of the covariance matrix of the observed signal to the temporal mean of the covariance matrix of the conventional trained estimate. We have also studied the performance of two Wiener like channel estimators based on our new estimate, and have compared their performances with that of a classical estimate of $\mathbf{\Gamma}$ used in the context of mono-user systems. The simulation results have shown that the new estimate outperforms quite significantly the classical one.

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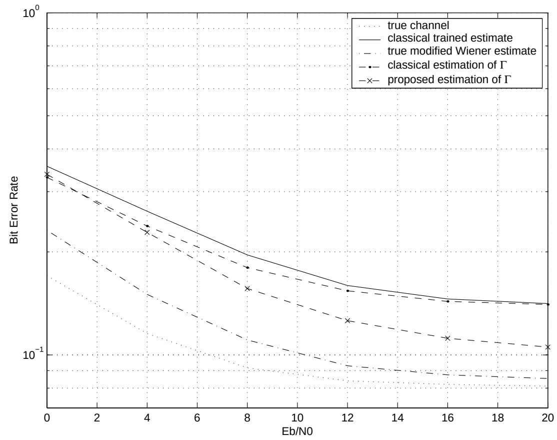

Fig. 1. Performance provided by modified Wiener estimates

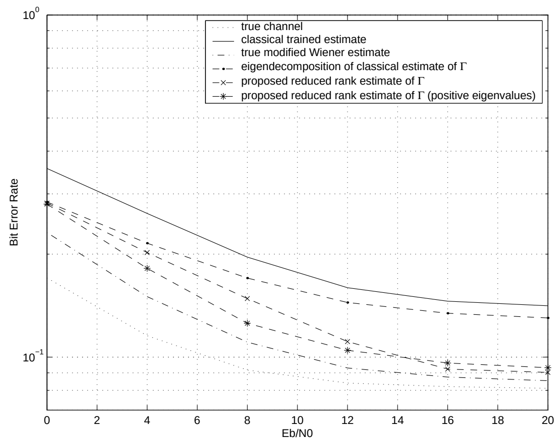

Fig. 2. Performance of reduced rank modified Wiener estimates