Improving the Performance of Blind CDMA 2D-RAKE Receivers with Phase Ambiguity in the Bit Decision Variable

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Abstract

Recently, a blind equalization algorithm for base station CDMA receiver exploiting antenna and path diversity, based on the principal eigenvector computation of a correlation matrix pair, has been presented [2]. The phase ambiguity in the bit decision variable obtained with this algorithm leads, for PSK, to the use of differential detection. In this article, we present an algorithmic modification that consists in estimating and cancelling the phase ambiguity in the bit decision variable, so that a more efficient coherent detector can be used. The gain in uplink capacity resulting from the use of this algorithmic modification is demonstrated via simulation.

1. Introduction

The CDMA 2D-RAKE single-user receiver exploits both the antenna and path diversity of the signals received at an antenna array to reduce the adverse effect of both multi-access interference (MAI) and fading [4]. The performance of this receiver is strongly dependent on the algorithm used to obtain the time-varying receiver coefficients required to coherently combined the various diversity branches [7]. The blind estimation of these coefficients is of particular interest for uplink reception since the use of a different pilot or training signal for each mobile would considerably reduce capacity.

A well known blind algorithm that can be used to estimate the optimal receiver coefficients is based on the computation of the principal generalized eigenvector of a correlation matrix pair [2], and is called the principal component (PC) algorithm.

There is an unavoidable sign ambiguity in the bit decision variable for any blind PSK equalization algorithm. Bits therefore have to be differentially encoded at the transmitter, and differentially decoded at the receiver. Furthermore, the use of a receiver coefficient estimation algorithm based solely on eigenvector computations, like the PC algorithm, results in the presence of a phase ambiguity in the bit decision variable, and leads to the use of differential bit detection. Assuming perfect phase estimation, the use of coherent bit detection followed by differential decoding of the detected bit sequence, instead of differential detection, would significantly reduce bit-error-rate (BER).

In this paper, we present a direct adaptation of the PC algorithm to balanced QPSK. We then propose to modify this algorithm by including a procedure which estimates and cancels the phase of the bit decision variable, thus allowing the subsequent use of a coherent bit detector instead of a differential one. The underlying assumption used in developing the new modified PC algorithm is that channel variations are slow compared to the bit duration, which is often the case in mobile communications. This assumption is frequently used in developing equalization schemes, and is, in any case, a condition for the proper operation of the PC algorithm without modifications. An estimation procedure can therefore be devised to track the bit decision variable phase required for the use of coherent bit detection. The proposed phase estimation procedure is based on the averaging of the last few bit decision variable phases, taking into account the phase ambiguity existing between successive receiver coefficient updates obtained via eigenvector computations. Note that although the modified PC algorithm developed in this paper is for balanced QPSK, the algorithm could easily be adapted to other types of modulation, like orthogonal M-ary, to allow the use of coherent detection.

The structure of the paper is as follows. The data model and the operation of the 2-D RAKE receiver are detailed in section 2. The PC algorithm adapted to QPSK is presented in the first part of section 3. In the second and last part of section 3, this PC algorithm is modified to include the newly proposed phase ambiguity canceler. Section 4 presents and discusses simulation realizations which demonstrate the gain in performance obtained by the use of our phase ambiguity canceler.

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2. Preliminary notions

2.1. Data Model

We consider an asynchronous CDMA system with multiple antennas at the base station only. The transmission system is modeled as a single-input multi-output (SIMO) discrete system, where the signals are sampled once every chip. Under the narrowband array assumption (the array size must be much smaller than the speed of light divided by the bandwidth of the incoming signal, B), the baseband model for the N_e-dimensional sampled received signal at the base station receivers is

$$\mathbf{s}(k) = \sum_{i=0}^{M-1} \mathbf{a}_i(k) z(k-i) + \mathbf{n}(k) \quad (1)$$

where M is the number of time-differentiable paths (TDP; in order for two paths to be time-differentiable, their relative delay of arrival must be greater then $1/B$ [9]), i is the TDP index, $\mathbf{a}_i(k)$ is the i^{th} complex path vector, whose N_e elements are called the channel coefficients, $z(k)$ is the chip sequence, and $\mathbf{n}(k)$ is the noise vector (including MAI). Assuming that each transmitted bit is spread into L chips, the aperiodic spreading code for the nth bit, $I(n)$, is denoted by $\mathbf{m}(n)=[m(nL) \ \dots \ m(nL+L-1)]^T$, where $|m(nL+k)|$ is normalized to one, so that $z(k) = I(n)m(k)$, with $n = \lfloor \frac{k}{L} \rfloor$. Note that balanced DQPSK is used, so that the spreading code is complex. The differentially coded transmitted bit, $I(n)$, is related to the information bit, $B(n)$, via $I(n)=B(n)I(n-1)$.

A vectorial formulation is obtained with the following definitions:

$$\mathbf{s}_N(k) \triangleq \left[\mathbf{s}^H(k) \quad \cdots \quad \mathbf{s}^H(k+N-1)\right]^H \quad (2)$$

$$\mathbf{z}_{MN}(k) \triangleq \left[z(k-M+1) \quad \cdots \quad z(k+N-1)\right]^T \quad (3)$$

$$\mathbf{n}_N(k) \triangleq \left[\mathbf{n}^H(k) \quad \cdots \quad \mathbf{n}^H(k+N-1)\right]^H, \quad (4)$$

with N a positive integer that will later be chosen to correspond to the order of the temporal filter used at each antenna element of the 2D-RAKE receiver. We then have

$$\mathbf{s}_N(k) = \mathbf{H}_{MN}(k)\mathbf{z}_{MN}(k) + \mathbf{n}_N(k). \quad (5)$$

with the matrix $\mathbf{H}_{MN}(k)$ characterized by the bloc elements

$$\{\mathbf{H}_{MN}(k)\}_{lm} = \mathbf{a}_{M-1+l-m}(k+l-1) \quad (6)$$

for $l=1,\dots,N$ and $m=1,\dots,N+M-1$, with $\mathbf{a}_i(k)=\mathbf{0}$ if $i > M-1$ or $i < 0$.

The following assumptions are made: the information symbols $I(n)=\pm 1$ are independent; the spreading codes are normalized complex binary, i.e. $m(k)=(\pm 1 \pm j)/\sqrt{2}$; the noise vector $\mathbf{n}_N(k)$ is zero-mean and white in time and space.

2.2. The 2D-RAKE Receiver

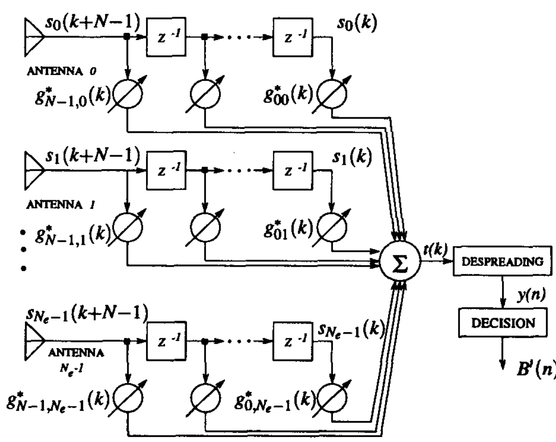

Figure 1. 2D-RAKE receiver for CDMA.

The standard RAKE receiver [5] is a scalar filter that combines the time-differentiable paths of a received signal. Recently, a space-time generalization of the RAKE filter has been proposed [1] to exploit not only the path diversity but also the spatial diversity of the channel. The 2D-RAKE receiver for CDMA is illustrated in Fig 1. In this figure, $s_j(k)$ is the signal received at the jth antenna and sampled at the chip rate; $g_{ij}(k)$ is the ith coefficient for the receiving filter of order N corresponding to the jth antenna; $t(k)$ is the output of the space-time filter, which is fed into the despreader; $y(n)$ is the despreader output, also called the decision variable; and $B'(n)$ is the information bit estimate. The despreader simply multiplies the spreading sequence for the nth bit with the filtered received signal corresponding to that same bit and does a summation over the chips. The signals being feed into the despreader are therefore at the chip rate while those coming out are at the bit rate. The decision as to which bit was transmitted is based on the despreader output.

In order for the 2D-RAKE receiver to properly detect the nth information bit $B(n)$, one must select the appropriate N_eN-dimensional channel-dependent weight vector $\mathbf{g}(k)=\left[g_{00}(k) \quad g_{01}(k) \quad \cdots \quad g_{N-1,N_e-1}(k)\right]^T$. One commonly used cost function to minimize is the chip-level mean-square error (MSE), defined as

$$E\left\{|t(k)-z(k)|^2\right\} = E\left\{|\mathbf{g}(k)^H\mathbf{s}_N(k)-I(n)m(k)|^2\right\}. \quad (7)$$

The optimal $\mathbf{g}(k)$ in that case is given by

$$\mathbf{g}(k) = \mathbf{R}^{-1}_{\mathbf{s}_N,\mathbf{s}_N}(k)\ \mathbf{r}_N(k), \quad (8)$$

where $\mathbf{R}_{\mathbf{s}_N,\mathbf{s}_N}(k)=E\left\{\mathbf{s}_N(k)\mathbf{s}_N^H(k)\right\}$ and $\mathbf{r}_N(k)= E\left\{\mathbf{s}_N(k)I(n)m^*(k)\right\}$. From (1)and (2), it can be shown that

$$\mathbf{r}_N(k) = \left[\mathbf{a}_0^H(k) \quad \mathbf{a}_1^H(k+1) \quad \cdots \quad \mathbf{a}_{N-1}^H(k+N-1)\right]^H, \quad (9)$$

where $\mathbf{a}_i(k)=\mathbf{0}$ if $i > M-1$, so that the optimal receiver coefficients can be obtained from (8) once the channel is estimated.

3. Proposed algorithms

In this section, we present two new algorithms that can be used to obtain the 2D-RAKE receiver coefficients when balanced DQPSK is used and the channel is time-varying.

3.1. Direct adaptation of the PC algorithm to balanced QPSK

This algorithm is based on the following observation [8]:

$$\mathbf{R}_{\mathbf{x}_N,\mathbf{x}_N}(n)-\mathbf{R}_{\mathbf{S}_N,\mathbf{S}_N}(n) \;=\; (L^2-L)\,\mathbf{r}_N(n)\,\mathbf{r}_N^H(n), \quad (10)$$

where

$$\mathbf{S}_N(n) \;=\; [\mathbf{s}_N(nL) \;\cdots\; \mathbf{s}_N(nL+L-1)], \quad (11)$$

$$\mathbf{x}_N(n) \;=\; \mathbf{S}_N(n)\mathbf{m}^*(n), \quad (12)$$

$\mathbf{R}_{\mathbf{x}_N,\mathbf{x}_N}(n) \triangleq E\{\mathbf{x}_N(n)\mathbf{x}_N^H(n)\}$, and $\mathbf{R}_{\mathbf{S}_N,\mathbf{S}_N}(n) \triangleq E\{\mathbf{S}_N(n)\mathbf{S}_N^H(n)\}$. This indicates that $\mathbf{r}_N(n)$ can be determined as the principal eigenvector of the difference between the post- and pre-despreading correlation matrices, $\mathbf{R}_{\mathbf{x}_N,\mathbf{x}_N}(n)$ and $\mathbf{R}_{\mathbf{S}_N,\mathbf{S}_N}(n)$ respectively. Once this eigenvector is calculated, the optimal receiver coefficients can be obtained from (8) with $\mathbf{R}_{\mathbf{s}_N,\mathbf{s}_N}(k)$ replaced by the estimated pre-despreading correlation matrix. Since the phase information is lost in the process of eigenvector computation, this type of approach leaves a phase ambiguity is the decision variable. Previous work on the subject have disregarded this phase ambiguity either because only the optimality of the array output signal to interference plus noise ratio, which is independent of phase, was considered, or because M-ary modulation with noncoherent detection was used. The algorithm presented in this paper is for PSK, so that one can not use noncoherent detection, and phase ambiguity can not be disregarded.

The adaptive PC algorithm is shown in Alg. 1. The phase ambiguity in the decision variable leads to the use of differential detection. More precisely, the differentially decoded information bit estimate $B'(n)$ is given by the sign of the real part of the product between the decision variable $y(n,0)=y(n)=\mathbf{g}^H(n)\mathbf{x}_N(n)$ and the conjugate of the decision variable $y(n,1)=\mathbf{g}^H(n)\mathbf{x}_N(n-1)$ corresponding to the previous time sample but obtained with the present receiver coefficients. The reason for using $y(n,1)=\mathbf{g}^H(n)\mathbf{x}_N(n-1)$ and not $y(n-1,0)=y(n-1)=\mathbf{g}^H(n-1)\mathbf{x}_N(n-1)$ in the differential detection process is that, although the channel is only slowly varying, there is not only a phase ambiguity in computing $\mathbf{g}(n)$, there is also a phase ambiguity *between* $\mathbf{g}(n)$ and $\mathbf{g}(n-1)$. This phase ambiguity can not be dealt with by a simple approach like normalizing $\mathbf{g}(n)$ so that a given element $g_{ij}(n)$ is always set to 1. This is due to the fact that the channel is time-varying, and the fading of the diversity branch corresponding to the element $g_{ij}(n)$ used as reference could result in a sudden phase change for that element, thus making the differential detection unreliable.

Initialize: select $\lambda \lesssim 1$
select δ, a small positive number
$\mathbf{R}_x(0) = \delta\mathbf{I}$
$\mathbf{R}_S(0) = \delta\mathbf{I}$
for $n = 0,1,\ldots$ **do**
$\mathbf{x}_N(n) = \mathbf{S}_N(n)\mathbf{m}^*(n)$
$\mathbf{R}_S(n+1) = \lambda\mathbf{R}_S(n) + \mathbf{S}_N(n)\mathbf{S}_N^H(n)$
$\mathbf{R}_x(n+1) = \lambda\mathbf{R}_x(n) + \mathbf{x}_N(n)\mathbf{x}_N^H(n)$
$\mathbf{a}(n) =$ the principal generalized eigenvector of the matrix pair $\{\mathbf{R}_S(n+1), \mathbf{R}_x(n+1)\}$
$\mathbf{g}(n) = [\mathbf{R}_S(n+1)]^{-1}\mathbf{a}(n)$
$y(n,0) = y(n) = \mathbf{g}^H(n)\mathbf{x}_N(n)$
$y(n,1) = \mathbf{g}^H(n)\mathbf{x}_N(n-1)$
$B'(n) = \text{sign}\{\text{real}(y(n,0)y^*(n,1))\}$
end for

Algorithme 1. Adaptation of PC to balanced QPSK

3.2. Modified PC algorithm for balanced QPSK

In this section, a modified version of the PC algorithm which uses coherent detection is presented. The motivation for trying to use coherent bit detection followed by a differential decoding of the detected bits, instead of direct differential detection, is that it reduces the BER provided that it is possible to obtain a sufficiently precise estimate of the phase of the decision variable $y(n)$. More precisely, assuming perfect phase estimation and a constant bit-energy-to-noise ratio E_b/N_0, if coherent bit detection followed by differential decoding of the detected bits is used, the BER is approximately equal to $\text{BER}_c=0.5\,\text{erfc}\sqrt{E_b/N_0}$ (worst case scenario, assuming the differentially encoded bit errors are not consecutive), while it is equal to $\text{BER}_d=0.5e^{-E_b/N_0}$ if differential detection is used. Fig. 2 illustrates the BER reduction encountered when the phase required for coherent detection is perfectly estimated, and coherent detection followed by differential decoding of detected bits is used instead of differential detection. More specifically the plotted function in Fig. 2 is $(\text{BER}_d-\text{BER}_c)/\text{BER}_d*100$ *versus* BER_d. This figure clearly demonstrates the BER improvement that can be achieved by using coherent detection provided the phase estimation is accurate.

It can be shown that [2]

$$\mathbf{x}_N(n) = I(n)L\mathbf{r}_N(n) + \sum_{k=0}^{L-1}\mathbf{i}_N(nL+k)m^*(nL+k) \quad (13)$$

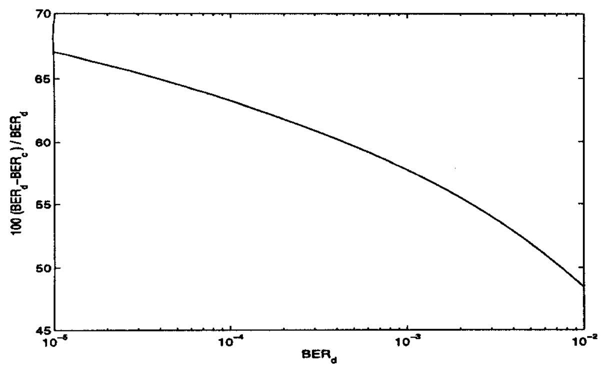

Figure 2. Percentage of BER reduction when using coherent detection followed by differential decoding of detected bits instead of differential detection, *versus* BER for differential detection.

where $\mathbf{i}_N(n) = \sum_{l=1, l\neq M}^{N+M-1} \mathbf{H}_{MN}(n)\mathbf{1}_l\mathbf{1}_l^H \mathbf{z}_{MN}(n) + \mathbf{n}_N(n)$ is an interference term which includes the effects of self interference and MAI, and $\mathbf{1}_l$ is a vector which only non-zero element is in the l^{th} position and is equal to one. It was shown via simulations in [3], that this interference term can be modeled as a zero-mean circularly complex Gaussian noise vector, spatially and temporally white, when the number of interfering users surrounding the base is high. The decision variables can therefore be modeled as

$$y(n,i) = \mathbf{g}^H(n)\mathbf{x}(n-i) = \pm I(n-i)r(n,i)e^{j\theta(n,i)} + \nu(n,i) \tag{14}$$

where $r(n,i)$ is the magnitude of the part of the decision variable which is associated with the desired signal, $\theta(n,i)$ is the corresponding phase, and $\nu(n,i)$ is an uncorrelated complex noise term due to interferences. Since the channel is only slowly varying, $\theta(n,i)$ will remain almost unchanged over a few bits i. We therefore have

$$E\left[\sum_{i=0}^{P-1\leq n} y^2(n,i)\right] \approx e^{j2\theta(n,0)} \sum_{i=0}^{P-1\leq n} r^2(n,i) + \sum_{i=0}^{P-1\leq n} E\left[\nu^2(n,i)\right], \tag{15}$$

where $E\left[\nu^2(n,i)\right] = 0$, and P is a positive integer which represents the number of bits over which the phase estimation is done. The phase estimate of $\theta(n) = \theta(n,0)$ can be obtained as

$$\theta'(n) = 0.5\angle\left\{\sum_{i=0}^{P-1\leq n} y^2(n,i)\right\}. \tag{16}$$

Once we have $\theta'(n)$, the phase of the decision variables $y(n,0)$ and $y(n,1)$ can be cancelled out so that they are *in phase* (or 180^o out of phase) with the transmitted bit $I(n)$, and coherent detection can be used.

The modified PC algorithm is shown in Alg. 2. Contrarily to the original algorithm, the differentially encoded bits are obtained by differential decoding of the coherently detected bits $I'(n,0)$ and $I'(n,1)$ instead of by direct differential detection. The parts of the standard PC and modified PC algorithms which differ are highlighted in Alg. 1 and 2.

Initialize: select $\lambda \lesssim 1$
select δ, a small positive number
select P, a positive integer
$\mathbf{R}_x(0) = \delta\mathbf{I}$
$\mathbf{R}_S(0) = \delta\mathbf{I}$
for $n = 0, 1, \ldots$ **do**
$\mathbf{x}_N(n) = \mathbf{S}_N(n)\mathbf{m}^*(n)$
$\mathbf{R}_S(n+1) = \lambda\mathbf{R}_S(n) + \mathbf{S}_N(n)\mathbf{S}_N^H(n)$
$\mathbf{R}_x(n+1) = \lambda\mathbf{R}_x(n) + \mathbf{x}_N(n)\mathbf{x}_N^H(n)$
$\mathbf{a}(n) =$ principal generalized eigenvector of the matrix pair $\{\mathbf{R}_S(n+1), \mathbf{R}_x(n+1)\}$
$\mathbf{g}(n) = [\mathbf{R}_S(n+1)]^{-1}\mathbf{a}(n)$
$y(n,0) = y(n) = \mathbf{g}^H(n)\mathbf{x}_N(n)$
$y(n,1) = \mathbf{g}^H(n)\mathbf{x}_N(n-1)$
$\theta'(n) = 0.5\angle\left\{\sum_{i=0}^{P-1\leq n}\{\mathbf{g}^H(n)\mathbf{x}_N(n-i)\}^2\right\}$
$I'(n,0) = \text{sign}\left[\text{real}\left(y(n,0)e^{-j\theta'(n)}\right)\right]$
$I'(n,1) = \text{sign}\left[\text{real}\left(y(n,1)e^{-j\theta'(n)}\right)\right]$
$B'(n) = I'(n,0)I'(n,1)$
end for

Algorithme 2. Modified PC for balanced QPSK

4. Simulation results and discussion

The performances of the PC algorithm with and without the newly proposed algorithmic modification are compared via simulations. We consider the reception at a base station with N_e=6 receiving antennas uniformly distributed around a horizontal circular array and separated by ten wavelengths. The synthetic baseband-equivalent antenna array received signals, for the user of interest, are generated by feeding a complex spread spectrum signal through the mobile vector channel simulator presented in [6]. Spatially and temporally Gaussian white noise is added to these signals to model MAI. The number of TDP used is M=4. Each time-differentiable path i is composed of a large number of time-indifferentiable subpaths lying in the same plane as the array and uniformly distributed in azimuth angle in $\theta_i \pm \Delta_i$, where θ_i is the mean angle of arrival and $2\Delta_i$ is the angle spread for the i^{th} path (see Table 1). The selected mobile speed is denoted by v, the carrier frequency f_c is $1GHz$ and the transmitted signal bandwidth B is $1.2288MHz$. The spreading factor is L=128. The voice activity factor is μ=0.4, and the number of interfering users is denoted by N_u.

The variance of the Gaussian white noises used to model MAI is given by μN_u times the average received power associated with the desired user. The order of the receiving filter at each antenna is equal to the number of TDP, i.e. $N = M = 4$.

	$i=0$	$i=1$	$i=2$	$i=3$
θ_i (degrees)	90	150	270	30
$2\Delta_i$ (degrees)	5	10	2	5

Table 1. Path angle of arrival parameters.

Fig. 3 illustrates the convergence behavior of the BER estimate *versus* time for v=15m/s (54km/h) and for v=30m/s (108km/h) with N_u=220 and P=15. The number of realizations used to obtained the estimate is 10000. We note in Fig. 3 that the proposed modification to the PC algorithm reduces the BER after convergence by about 45% if v=15m/s and by about 20% if v=30m/s. Although not shown here, similar curves have been obtained for v=15m/s and N_u equal to 200, 240, 260 and 300, and the corresponding reductions in BER after convergence are 50%, 45%, 40% and 29%.

The BER reduction ultimately translates into a more reliable transmission or into an increase in capacity. To help visualize the potential increase in capacity, Fig. 4 illustrates the BER estimate after convergence of the PC algorithm with and without the modification, *versus* the number of co-channel interferes (N_u), with v=15m/s. For a required maximum BER of 0.01, the maximum number of co-channel interferes goes from 233 to 273 if the modified algorithm is used instead of the standard PC algorithm, an increase of 17%.

From these observations and other simulation results obtained for different antenna configurations, number of co-channel interferes N_u, and mobile speed $v \leq 30m/s$, we conclude that our modified algorithm significantly outperforms the standard PC algorithm under practical operating conditions.

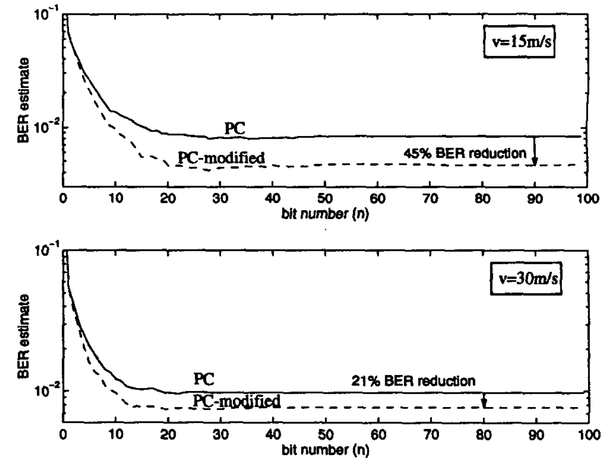

Figure 3. BER estimate *versus* bit index for N_u=220, v=15m/s and v=30m/s, obtained with PC and PC-modified.

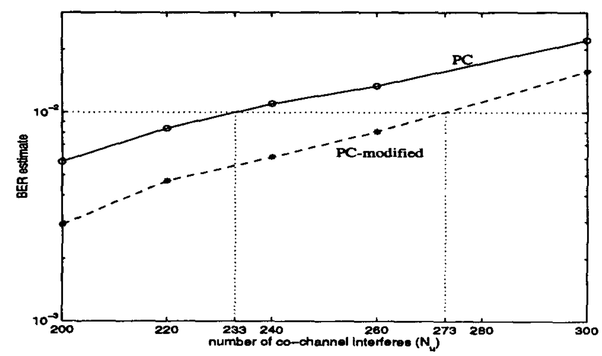

Figure 4. BER estimate after convergence *versus* the number of co-channel interferes (N_u) for v=15m/s obtained with PC and PC-modified.

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