Performance Evaluation of Space-Time Block Coding Using a Realistic Mobile Radio Channel Model *

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Abstract

This paper presents a performance evaluation of Space-Time Block Coding (STBC) employing a realistic mobile radio channel model in macrocellular and urban environment. The Bit Error Rate (BER) is computed by Monte-Carlo simulations in the down-link to evaluate its sensibility to channel correlation. We consider a horizontal uniform linear array at the Base Station (BS) formed by up to four antenna elements, and one and two uncorrelated antenna elements at the Mobile Station (MS). The channel model includes the probability density function (pdf) of the azimuth and delay of the impinging waves and their expected power conditioned on the azimuth and delay. The statistical properties of the model are extracted from macrocellular measurements made in urban environments. Simulation results show that the use of STBC can provide significant gains with acceptable sensibility to the channel correlation under realistic conditions.

1. Introduction

The demand for wireless communications is experiencing unprecedented growth. Next generation systems must support many users at high data rate under difficult propagation conditions and therefore require development of more efficient communications techniques. Spatial filtering and advanced diversity schemes, combining antenna array with signal processing in both space and time, are some of the potential techniques which can significantly increase the capacity of terrestrial cellular systems.

Recent information theory research has shown that the rich-scattering wireless channel is capable of enormous theoretical capacities if the multipath is properly exploited. Multiple-element transmit and receive antennas have shown very promising results for improved bit error rate and spectral efficiency [1]. Transmit diversity has been studied extensively as a method to combat fading in wireless channels because of its relative simplicity of implementation and feasibility of exploiting multiple antennas at the base station.

Space-time trellis coding (STTC) and Space-time block coding (STBC) have been proposed which combine signal processing at the receiver with coding techniques appropriate to multiple transmit antennas providing significant gains. STTC perform extremely well when the number of transmit antennas is fixed [2]. The decoding complexity of STTC increases exponentially as a function of both the diversity level and the transmit rate. On the other hand, STBC [3], developed on the base of a simple scheme for transmission using two transmit antennas [4], generalize the transmission scheme to an arbitrary number of transmit antennas and are able to fully achieve their inherent diversity [5]. These coding techniques retain the property of having a very simple maximum likelihood decoding algorithm, based only on linear processing at the receiver, because orthogonal design allows signal decoupling. This scheme does not require any bandwidth expansion or feedback from the receiver to the transmitter and its computational complexity is similar to that of Maximal-Ratio Receiver Combining (MRRC).

Space-time coding has been studied under ideal propagation conditions, but prior to its application into real systems, it is still necessary to evaluate its sensibility with respect to critical parameters and system features, real environments and applications.

STBC has been included as an option in the air interface specifications of third generation mobile communications. UTRA considers the simplest case, which consists on block coding for two transmit antennas having a QPSK constellation [6]. Existing publications about STBC performance evaluation consider mostly rich-scattering

* This work was partially supported by the European Commission under IST Project METRA; the Spanish Government (CYCIT) TIC98-0412 TIC98-0703 TIC99-0849; and the Catalan Government (CIRIT) 1998SGR 00081

wireless channel models, with completely uncorrelated signals at each antenna. Such situation occurs where the azimuth spread is sufficiently large so that channels observed from each antenna are completely independent. The required separation depends on the angular spread, that is the angle over which the signal arrives at the receiver. In mobile units (MS), which are generally surrounded by reflectors, the angular spread is typically 360°, and quarter-wavelength spacing of the antennas is sufficient. In outdoor systems with high base station antennas, located above the clutter, the angular spread may be only a few degrees, and a horizontal separation of 10 to 20 wavelengths is required.

In this work, BER evaluations of STBC by Monte-Carlo simulations are performed considering a horizontal uniform linear array (ULA) at the BS, in macro cellular and urban environments. A stochastic spatial radio channel model is used, which includes both the delay and azimuth dispersion [7, 8]. Probability distributions of the model are extracted from experimental data collected during extensive measurement campaigns.

Antenna elements at the MS are supposed to receive uncorrelated signals, whereas at the BS the mutual correlation between elements is given implicitly by the physical considerations of the channel model. The correlation coefficient between two elements at the BS is considered as a very important figure of merit to measure the performance sensibility gain of the transmission system under study.

We consider a wireless communication system with N antenna elements at the BS (transmitter) and M antenna elements at the MS (receiver). At each time slot, signals are transmitted simultaneously from the N antennas. The channel is assumed to be quasi-static and flat, so that the path gains are constant over each transmitted symbol block, and vary from one to another. Perfect state channel information is assumed available at the receiver.

2. Channel Model

In multipath radio channel, multiples replicas of the transmitted signal are received in general from different directions seen from BS, with different time delays, as shown in Figure 1, where an arbitrary antenna array at the BS and one antenna element at the MS is considered.

The received baseband signal vector at the BS antenna array can be expressed as

$$\mathbf{Y}(t) = \sum_{l=1}^{L} \alpha_l \mathbf{c}(\varphi_l) u(t-\tau_l) + \mathbf{N}(t) \qquad (1)$$

The components of $\mathbf{Y}(t) = [Y_1(t),...,Y_N(t)]^T$ are the output of the N antenna elements and $u(t)$ is the transmitted signal. The parameters α_l, τ_l and φ_l are the complex amplitude, delay and incidence azimuth of the lth impinging wave, respectively.

For an ULA on the y axis, the components of the steering vector $\mathbf{c}(\varphi) = [c_1(\varphi),...,c_N(\varphi)]^T$ can be expressed as $c_m(\varphi) = f_m(\varphi)\exp[-(m-1)2\pi d/\lambda sin(\varphi)]$, where $f_m(\varphi)$ is the complex field pattern of the mth array element, d is the element spacing and λ is the wavelength. The components of the noise vector $\mathbf{N}(t) = [N_1(t),...,N_N(t)]^T$ are asummed to be independent complex white Gaussian processes with identical power density. P_l represents the expected power of the lth wave.

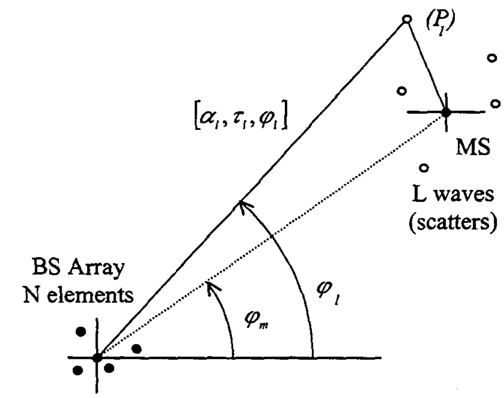

Fig. 1.- Channel model reference.

We assume that the random vectors $[\alpha_l, \tau_l, \varphi_l]$, for $l=1:L$, are independent and identically distributed.

If the instantaneous power azimuth-delay spectrum is defined as [7, 8]

$$P_i(\varphi,\tau) = \sum_{l=1}^{L} |\alpha_l|^2 \delta(\varphi - \varphi_l, \tau - \tau_l) \qquad (2)$$

the power azimuth-delay spectrum is

$$P(\varphi,\tau) = E\{P_i(\varphi,\tau)\} \qquad (3)$$

The power azimuth spectrum (PAS) and power delay spectrum (PDS) are derived from

$$P_A(\varphi) = \int P(\varphi,\tau)d\tau \qquad (4)$$

and

$$P_D(\tau) = \int P(\varphi,\tau)d\varphi \qquad (5)$$

If the joint pdf of the azimuth and delay is $f(\varphi,\tau)$, their marginal pdf's are

$$f_A(\varphi) = \int f(\varphi,\tau)d\tau \qquad (6)$$

and

$$f_D(\varphi) = \int f(\varphi,\tau) d\varphi \quad (7)$$

Then, the unconditional power spectrum is derived as

$$P(\varphi,\tau) \propto E\{|\alpha_l|^2 \mid \varphi,\tau\} f(\varphi,\tau) \quad (8)$$

where, the expected power of the waves conditioned on their azimuth and delay is

$$E\{|\alpha|^2 \mid \varphi,\tau\} = E\{|\alpha_l|^2 \mid \varphi_l = \varphi, \tau_l = \tau\} \quad (9)$$

The PAS and PDS can be expressed as

$$P_A(\varphi) \propto E\{|\alpha|^2 \mid \varphi\} f_A(\varphi) \quad (10)$$

and

$$P_D(\tau) \propto E\{|\alpha|^2 \mid \tau\} f_D(\tau) \quad (11)$$

where $E\{|\alpha|^2 \mid \varphi\}$ and $E\{|\alpha|^2 \mid \tau\}$ are the expected powers of the waves conditioned on their azimuth an delay, respectively.

The realistic model considered here, determines experimentally the functions $P(\varphi,\tau)$, $P_A(\varphi)$, $P_D(\tau)$, $f_A(\varphi)$, $f_D(\tau)$, $E\{|\alpha|^2 \mid \varphi\}$ and $E\{|\alpha|^2 \mid \tau\}$ [7, 8].

Estimates of $P(\varphi,\tau)$ are obtained by averaging $P_l(\varphi,\tau)$ over a distance of 100λ.

From experimental measurements, it is observed that Laplacian function matches the estimates PAS, as

$$P_A(\varphi) \propto \exp(-\sqrt{2}\frac{|\varphi|}{\sigma_A}), \text{ for } \varphi \in [-180°;+180°] \quad (12)$$

and exponential decaying functions matches the PDS as

$$P_D(\tau) \propto \begin{cases} \exp(-\frac{\tau}{\sigma_D}), & \text{for } \tau > 0 \\ 0 & \text{otherwise} \end{cases} \quad (13)$$

where σ_A and σ_D are defined as the azimuth and delay spread, and correspond to the root second central moment of $P_A(\varphi)$ and $P_D(\tau)$, respectively.

For the pdf it is found that a Gaussian function provides a good match for the estimated azimuth, as

$$f_A(\varphi) \propto \exp(-\frac{\varphi^2}{2\sigma_A^{*}}) \quad (14)$$

where σ_A^{*} is the standard deviation computed for φ. (φ respect to φ_m)

For the estimated pdf of delay, it is found that an exponential decaying function is appropriate as

$$f_D(\tau) \propto \exp(-\frac{|\tau|}{\sigma_D^{*}}) \quad (15)$$

where σ_D^{*} is the standard deviation computed for τ.

As a consequence of the model, the expected power conditioned on azimuth can be expressed as

$$E\{|\alpha|^2 \mid \varphi\} \propto \exp(\frac{\varphi^2}{2\sigma_A^{*}} - \sqrt{2}\frac{|\varphi|}{\sigma_A}) \quad (16)$$

and the estimated expected power conditioned on delay can be expressed as

$$E\{|\alpha|^2 \mid \tau\} \propto \exp(-\tau(\frac{1}{\sigma_D} - \frac{1}{\sigma_D^{*}})) \quad (17)$$

The expected power conditioned on azimuth exhibits two local minimums at $|\varphi_{min}| = \sqrt{2}\sigma_A^{*} / \sigma_A$, increasing rapidly as φ increases.

From a physical point view, the power can not suddenly increasing with $|\varphi|$, for this reason, for implementing of radio channel simulator, it is suggested to truncate (16) so that $E\{|\alpha|^2 \mid \varphi\} = E\{|\alpha|^2 \mid \varphi_{min}\}$, for $|\varphi| > |\varphi_{min}|$.

It has been experimentally verified that the conditional distribution functions $f_A(\varphi)$ and $f_D(\tau)$ are essentially independent of τ and φ therefore, it is suggested the following decomposition for the joint azimuth-delay pdf

$$f(\varphi,\tau) = f_A(\varphi) f_D(\tau) \quad (18)$$

Finally, the channel can be simulated with an expected power function of the *l*yh wave (P_l) given by

$$P_l(\varphi,\tau) \propto \begin{cases} \exp(\frac{\varphi^2}{2\sigma_A^{*}} - \sqrt{2}\frac{|\varphi|}{\sigma_A}) \exp(-\tau(\frac{1}{\sigma_D} - \frac{1}{\sigma_D^{*}})) & \text{for } |\varphi| \le \varphi_{min} \\ \exp(-\frac{\sigma_A^{*}}{\sigma_A}) \exp(-\tau(\frac{1}{\sigma_D} - \frac{1}{\sigma_D^{*}})) & \text{otherwise} \end{cases} \quad (19)$$

where, from experimental results and for a typical urban environment with high BS antenna position, it is found that the $\sigma_A^{*} \approx 1.38\sigma_A$ and $\sigma_D^{*} \approx 1.17\sigma_D$.

3. Space-Time Block Coding

STBC, developed on the base of a simple scheme for transmission using two transmit antennas, generalize the transmission scheme to an arbitrary number of transmit antennas [4] and are able to achieve significant diversity gains [3]. This coding strategy retain the property of having a very simple maximum likelihood decoding algorithm, based only on linear processing at the receiver, because orthogonal design allows signal decoupling. This scheme does not require any bandwidth expansion or feedback from the receiver to the transmitter and its computation complexity is similar to that of Maximal-Ratio Receiver Combining (MRRC).

The use of complex constellations in STBC permits to increase the spectral efficiency. By doing this, coding matrices use different constellation as N changes [3].

STBC are constructed by a P x N transmission matrix, whose entries are linear combinations of the signal constellation employed at baseband. N streams of P symbols are transmitted simultaneously from each antenna element. The received signal at each of M receiver antennas is a linear superposition of the N transmitted signals, perturbed by noise.

The basic scheme of STBC, that has been proposed as a transmission diversity option for the 3th Generation Systems, UTRAN, consists on the use of two transmit antennas at the BS [6].

Consider N=2, and M antenna elements at the MS, the code matrix for this case is

$$\mathbf{G}_2 = [\mathbf{C}_1 \quad \mathbf{C}_2], \text{ with } \mathbf{C}_1 = \begin{bmatrix} X_1 \\ -X_2^* \end{bmatrix} \text{ and } \mathbf{C}_2 = \begin{bmatrix} X_2 \\ X_1^* \end{bmatrix}$$

Thus, the symbol sequence to be transmitted by an antenna is $X_1, -X_2^*$, and by the other X_2, X_1^*. Orthogonality is obtained since $\mathbf{C}_1^H * \mathbf{C}_2 = 0$.

Assuming flat fading, i.e. no time dispersion, the complex baseband signals received by the mth antenna element at the MS, at time t_1 and t_2 (P=2), is

$$r_m(t_1) = \alpha_{1,m} \cdot X_1 + \alpha_{2,m} \cdot X_2 + n_m(t_1) \tag{20}$$

and

$$r_m(t_2) = -\alpha_{1,m} \cdot X_2^* + \alpha_{2,m} \cdot X_1^* + n_m(t_2) \tag{21}$$

where $\alpha_{n,m}$ represents the complex channel gain from the nth transmit antenna to the mth receive antenna, assumed constant during the transmission of a symbol block. The term $n_m(t)$ represents the Gaussian complex baseband noise received by the mth antenna element. The channel characteristics are assumed known to the receiver

In this channel model, the complex channel gain corresponds to the total contribution of the L waves arriving from the nth transmit antenna to the mth receive antenna. The local scatteres at the MS are assumed to provide uncorrelated signals for each antenna element.

The decoding algorithm selects the symbols that minimize the metric

$$\Psi = \sum_{m=1}^{M} \Big(|r_m(t_1) - \alpha_{1,m} \cdot X_1 - \alpha_{2,m} \cdot X_2|^2 + |r_m(t_2) + \alpha_{1,m} \cdot X_2^* - \alpha_{2,m} \cdot X_1^*|^2 \Big) \tag{22}$$

which, due to orthogonality, can be decomposed as

$$\Psi_1 = \left| \left[\sum_{m=1}^{M} \big(r_m(t_1) \cdot \alpha_{1,m}^* + r_m^*(t_2) \cdot \alpha_{2,m} \big) \right] - X_1 \right|^2 + \left(-1 + \sum_{m=1}^{M} \sum_{n=1}^{2} |\alpha_{n,m}|^2 \right) \cdot |X_1|^2 \tag{23}$$

and

$$\Psi_2 = \left| \left[\sum_{m=1}^{M} \big(r_m(t_1) \cdot \alpha_{2,m}^* - r_m^*(t_2) \cdot \alpha_{1,m} \big) \right] - X_2 \right|^2 + \left(-1 + \sum_{m=1}^{M} \sum_{n=1}^{2} |\alpha_{n,m}|^2 \right) \cdot |X_2|^2 \tag{24}$$

The following equation represents the signal vector received by each antenna element, for M=1

$$\mathbf{r} = \begin{bmatrix} r(t_1) \\ r^*(t_2) \end{bmatrix} = \mathbf{H} \cdot \begin{bmatrix} X_1 \\ X_2 \end{bmatrix} + \begin{bmatrix} n(t_1) \\ n(t_1) \end{bmatrix} \tag{25}$$

the channel matrix

$$\mathbf{H} = \begin{bmatrix} \alpha_1 & \alpha_2 \\ \alpha_2^* & -\alpha_1^* \end{bmatrix} \tag{26}$$

has orthogonal columns. This shows symbols X_1 and X_2 to be sent through two orthogonal vector channels. This is the reason why use of a 2-branch STBC provides full-rate transmission with two levels of diversity [9].

The other coding matrices considered in this work, for N=3 ($\mathbf{G}_3$) and N=4 ($\mathbf{G}_4$), are the following

$$\mathbf{G}_3 = \begin{bmatrix} X_1 & X_2 & X_3 \\ -X_2 & X_1 & -X_4 \\ -X_3 & X_4 & X_1 \\ -X_4 & -X_3 & X_2 \\ X_1^* & X_2^* & X_3^* \\ -X_2^* & X_1^* & -X_4^* \\ -X_3^* & X_4^* & X_1^* \\ -X_4^* & -X_3^* & X_2^* \end{bmatrix} \qquad \mathbf{G}_4 = \begin{bmatrix} X_1 & X_2 & X_3 & X_4 \\ -X_2 & X_1 & -X_4 & X_3 \\ -X_3 & X_4 & X_1 & -X_2 \\ -X_4 & -X_3 & X_2 & X_1 \\ X_1^* & X_2^* & X_3^* & X_4^* \\ -X_2^* & X_1^* & -X_4^* & X_3^* \\ -X_3^* & X_4^* & X_1^* & -X_2^* \\ -X_4^* & -X_3^* & X_2^* & X_1^* \end{bmatrix}$$

Their decoding metrics decomposition can be found in [3].

4. Simulations Results

In this section we compare the effect of channel correlation at the BS end on coding gains using different N and M combinations (referred as NxM). Mainly a transmission of 1[b/s/Hz] is considered. Additionally, to illustrate the effect of the channel correlation in a 2x2-QPSK scheme (2[b/s/Hz], as in UTRA), the BER performance evaluation is also included for this case.

The number of transmit and receive antenna is set to 2, 3 and 4, and 1 and 2, respectively (N, M). For N=2, a BPSK constellation is used, whereas for N=3 and 4, a QPSK constellation is used.

A uniform linear array of N antenna elements is consider at the BS, whose relative spacing elements is d/λ, and the mean arrival angle from the MS is φ_m, respect to the broadside direction. Since only flat fading is taken account, the channel model considers only angular dispersion in order to compute the expected power of the waves (23).

The number of waves (L) is modeled by a Poisson process with an expected value of 25 paths. The instantaneous path gains of the waves (α_l), are determined according to a Rayleigh process. The expected power and path gain are normalized so that the total power received by the *m*th antenna is 1. The power of the Gaussian complex noise is set according to the required Signal to Noise Ratio (SNR).

In order to generate M uncorrelated signals at the MS (for the M antenna elements), each of the M sets of N coefficients are independently generated using the channel model.

Figure 2 shows the envelope correlation of the channel gain ($\alpha_{n,m}$), computed for several combinations of the relative spacing between antenna elements, mean angle of arrival and the angular spread at the BS. From these results, it is found that the correlation coefficient (CC) can be approximately represented by a function of the variable δ, which involves the product of the projected spacing and the angular deviation. Given an angular deviation and a mean angle of arrival, a minimum relative spacing is required to limit the correlation coefficient.

Figures 3 and 4 show the BER computed for a 2x2 and 4x2 STBC schemes, respectively, both for 1[b/s/Hz] and for different correlation coefficient (CC) values.

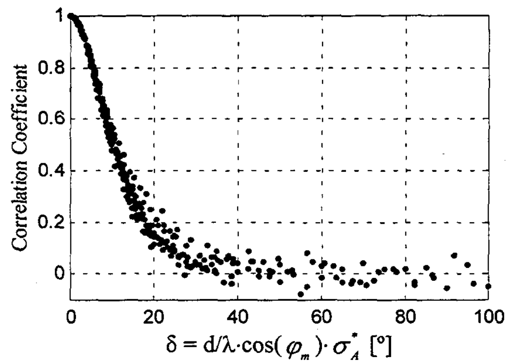

Fig. 2.- Correlation Coefficient for BS antenna elements.

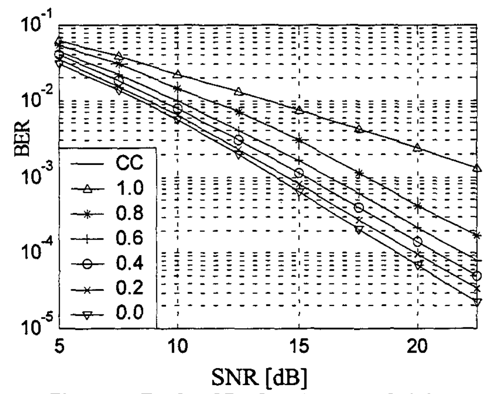

Fig. 3.- BER for STBC, 2Tx 1Rx, 1[b/s/Hz].

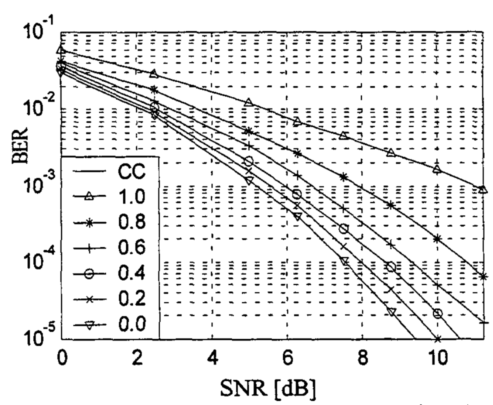

Fig. 4.- BER for STBC, 4Tx 2Rx, 1[b/s/Hz].

Figures 5 and 6 show the SNR required for BER=10^{-4} and BER=10^{-4}, as a function of CC, and for different combinations of NxM schemes, respectively.

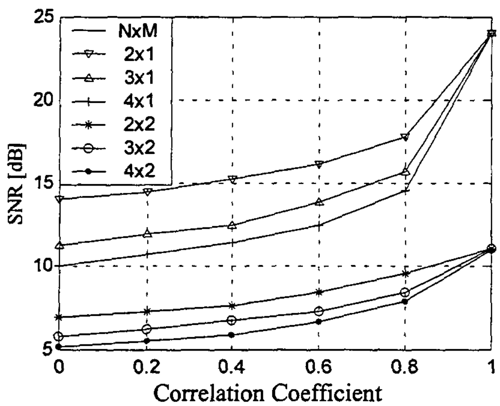

Fig. 5.- SNR for BER=10^{-3}, 1[b/s/Hz].

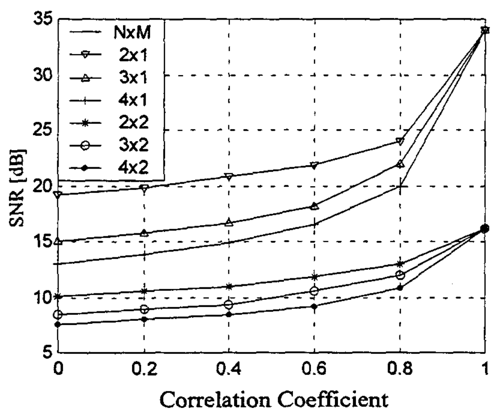

Fig. 6.- SNR for BER=10^{-4}, 1[b/s/Hz].

Figures 7 and 8 show the coding gain loss as a function of CC (ideal case CC=0).

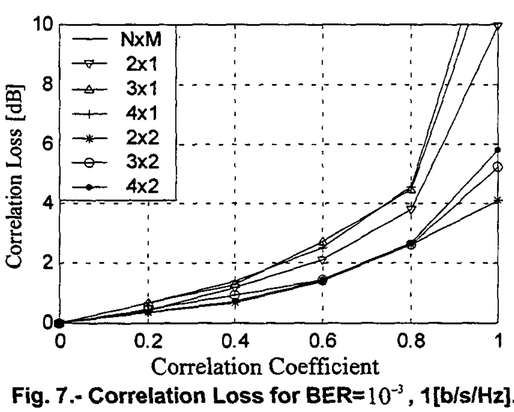

Fig. 7.- Correlation Loss for BER=10^{-3}, 1[b/s/Hz].

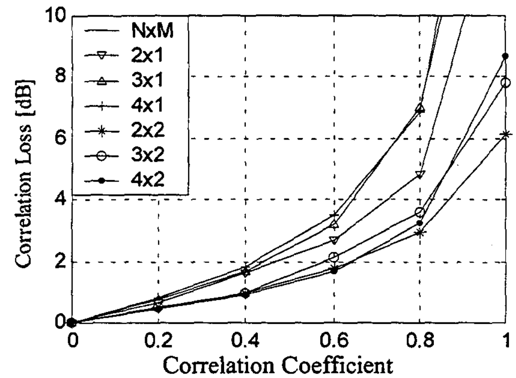

Fig. 8.- Correlation Loss for BER=10^{-4}, 1[b/s/Hz].

From these results, it can be observed that in general the sensibility to the channel correlation is mildly linear (in dB) until CC≈0.5, and increases rapidly as CC increase from this value. Also a minor sensibility is observed as the number of receive antennas (M) increases, as well as for a higher BER reference.

In all cases it is observed that when CC=1 performance corresponds to the MRRC with only one transmit antenna, showing that there is no transmission diversity. When the correlation coefficient is 0, maximum diversity is achieved.

It can also be inferred that for a maximum CC of around 0.4, the effect of the channel correlation is not significant. This implies for example that for BER=10^{-4} of a 4x2 scheme, with a typical angular spread of 6°, and a covered sector of 120° by the BS, a minimum relative spacing of around 4λ is required. Consequently, it is estimated that this restriction for the maximum CC is a realistic condition to achieve.

Figure 9 shows the BER performance for a 2x2 scheme using a QPSK constellation to transmit 2[b/s/Hz]. For comparison, curves for 1x1 QPSK and 2x2 QPSK (CC=0) are included. Considering the feasibility of STBC used in the up-link, BER is computed also changing the correlation channel to the receive end. Results are the same for down-link as well as for up-link.

Because of the change from BPSK to QPSK, the BER performance of a 2x2 - 2[b/s/Hz] scheme is equivalent to a 2x2 - 1[b/s/Hz] scheme, with an increment of 3 dB in the SNR required. And consequently, the same sensibility to the correlation coefficient is obtained.

Equivalent performance of down-link and up-link occurs because the tendency of a 2x2 scheme to a 2x1 scheme, when CC tends to 1, is compensated by the array gain at the BS. The special case of using 2-branch STBC is also interesting because it is possible to combine beam forming and polarization diversity.

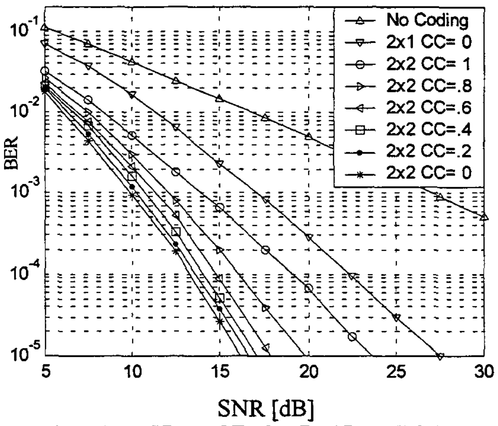

Fig. 9.- BER for STBC, 2Tx 2Rx, 2[b/s/Hz].

5. Conclusions

In this paper the effect of the channel correlation on the BER performance of STBC is studied. A realistic channel model is used to evaluate the sensibility to the correlation coefficient in real conditions of an urban macrocellular environment.

An horizontal uniform linear antenna array is consider at the BS, whereas an arbitrary antenna array of uncorrelated elements is consider at the MS.

The correlation coefficient (CC) can be expressed as a function of antenna element spacing, direction of arrival and angular spread at the BS.

The BER performance degradation due to the correlation in the Multiple Input - Multiple Output (MIMO) channel, was analyzed and showed that the degradation is not significant if the correlation coefficient remains under the value of 0.4. Consequently, use of STBC allows significant gains under realistic channel conditions.

Finally, it is also illustrated that, for fully correlated channels, the performance corresponds to the MRRC with only one transmit antenna, verifying that there is no transmission diversity.

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