A new Criterion for Transmit Diversity and Beamforming in Mobile Communications

Danilo Zanatta Filho and Luc Féty

***Abstract*— We consider a wireless communication system in which the base station is equipped with an antenna array and the mobile user has a single antenna. Classically, in this configuration, the antenna array at the base station is used to perform either beamforming or transmit diversity. We propose a scheme that exploits the antenna array to perform beamforming in order to maximize the diversity of the channel seen at the receive end, i.e., the mobile user. This scheme is based on the criterion of minimizing the variation of the power received by the mobile user due to fading. Moreover, we propose an adaptive method to optimize this criterion, which is a type of CM (Constant Modulus) criterion.**

***Keywords*— Adaptive antennas, smart antennas, downlink beamforming, transmit beamforming, transmit diversity, downlink diversity, CM, Constant Modulus, CMA, Constant Modulus Algorithm, mobile wireless systems.**

I. INTRODUCTION

The evolution of the wireless communication systems towards 3GB (3rd Generation and beyond) is propelled by Internet access and increasing demand for data-based services. Most of these services, such as internet surfing, are downlink-intensive, in opposition to voice services, which demand the same data rate in both directions. These factors lead to an increasing demand for higher system capacity and higher data rates, mostly in the downlink, which can be achieved by a better link quality in terms of higher decoding SNR (Signal to Noise Ratio).

The wireless propagation channel is formed by multiple paths that sum with each other to construct the channel. This sum of multiple paths leads to a fluctuation of the received power in time, which is known as *fading*. Poor performance due to prolonged deep fading of the channel is one of the main problems faced when using the wireless channel.

In order to counteract deep fading, one can benefits from the decorrelation of the channel to combine uncorrelated copies of the same signal. It is very unlikely that all these copies are in a fade simultaneously. Hence, the chance that a deep fading occurs is greatly reduced and, when it occurs, its duration is also reduced when compared to the original scenario of only one copy of the signal. This strategy is called *diversity* and it may be the single most important contributor to reliable wireless communications. Increasing the number of uncorrelated copies of the same signal has the effect of reducing the probability that the combined signal is in a fade. The number of possible copies is called the *diversity order* of the channel.

It is well established that the use of multiple antennas can improve the performance of a wireless communication system in a fading environment [1], [2]. From the point of view of diversity, the use of multiple antennas increases the diversity order due to spatial uncorrelated copies of the signal, leading to a diversity order greater than one even for flat fading channels. Although multiple antennas may be employed at either the base station (BS), mobile unit (MU), or both, it is most cost effective and practical to employ multiple antennas at the base station only. Hence, this article is restricted to the case of employing multiple antennas at the BS.

The considered case of employing multiple antennas at the BS only has already been contemplated in the literature for receive diversity (RD) in the uplink [2]. The idea is to combine the antennas outputs according to some criterion in order to take advantage of the spatial diversity of the received signal. This is called receive diversity since the diversity is exploited at the receiver. This is a well studied problem in the literature and the main drawbacks to practical implementation are the required computational cost and, mostly, hardware implementation issues.

However, the more challenging problem of profiting from the diversity in the downlink is yet an open problem. Recent works propose to use some processing prior to transmission by the multiple antennas in order to achieve a diversity gain at the MU receiver [3]. Since the processing is done at the transmission side, this scheme is called transmit diversity (TD). With the use of TD, the MU receiver is kept simple (only one antenna) and yet profits from diversity gain. The main theoretical impairment for TD is that the channel is unknown at transmission time.

Among the different schemes for TD proposed in the literature, we can identify two main categories: one based on the knowledge of the downlink channel and another which does not require the knowledge of the downlink channel. The main idea behind the TD scheme without channel knowledge is that each antenna transmit an uncorrelated version of the desired signal. The receiver (MU) can hence combine these different versions of the desired signal in order to profit from the spatial diversity of the multipath channel. Since each antenna is used regardless of the others, the radiation is omnidirectional and it is not possible to perform beamforming. In a multiuser context, this approach has the weakness of generating a high interference level due to omnidirectional radiation. Examples of these schemes, called *open loop TD*, are the delay diversity schemes [4], [5] and space-time codes [6], with the Alamouti scheme [7] being one particular case.

Danilo Zanatta Filho is with the DSPCom-DECOM-FEEC, Universidade Estadual de Campinas (UNICAMP), Brazil and the Laboratoire d'Électronique et Communication, Conservatoire National des Arts et Métiers (CNAM), Paris, France. Luc Féty is with the Laboratoire d'Électronique et Communication, Conservatoire National des Arts et Métiers (CNAM), Paris, France. E-mails: daniloz@decom.fee.unicamp.br and fety@cnam.fr

The other approach, which requires the knowledge of the downlink channel, is based on the estimation of this channel by the MU. This information is then fed back to the BS, where the channel estimation is used to carry out TD. Switched transmit diversity [8] and transmit adaptive array [9] are examples of feedback-based TD, also called *close loop TD*. The advantage of using feedback is that the BS can have an estimation of the impulse response of the downlink channel. This knowledge makes possible to the BS to compensate for the phases of each multipath, eliminating thus the fading. There are two drawbacks, however: useful rate reduction and feedback delay. The insertion of training symbols is needed so that channel estimation is possible at the MU. This estimation must be then fed back to the BS using the uplink channel. Although both points imply on useful-rate reduction in both the downlink and uplink, the impact of this rate reduction on the system is not remarkable and it is worthy due to the corresponding gain. On the other hand, the feedback delay, which is the delay between the downlink channel estimation at the MU and its exploitation by the BS, is a more critical issue. This delay makes the feedback approach not very robust in practical conditions since the downlink channel estimation is very likely to be out-of-date at the moment of its use by the BS. Indeed, it only takes the displacement of a fraction of wavelength to cause a significant change of the multipath phases, leading to a completely different fading condition. In this case, trying to compensate for the multipath phases is not possible and may even create a deep fading instead of a milder one.

In this work, we propose a diversity technique that is directly based on the criterion of minimizing the variation of the power received by the MU. This technique uses the second order statistics of the downlink channel. The channel covariance matrix (CCM) is formed from the physical characteristics of the channel (the physical paths), which does not change as quickly as the multipath phases. Hence, we can rely on this relatively stationary information to perform TD. Moreover, the CCM is of much interest since the downlink channel covariance matrix (DCCM) can be estimated directly at the BS, avoiding feedback. For TDD (time division duplex) systems, this is a very straightforward task. Since both links use the same frequency, the uplink channel covariance matrix (UCCM) is identical to the DCCM, and the estimation of the former can be easily obtained form the received signal in the uplink. Although in FDD (frequency division duplex) systems the multiple antenna response (i.e., the steering vector) differs from one link to the other due to their frequency separation, it is possible to estimate the DCCM from the UCCM, as showed by Asté in [10], [11].

Having estimated the DCCM, several works (see for example [12]–[14]) proposes to use this information to perform beamforming by using a purely spatial filter. We propose, however, to use a space-time filter to maximize the diversity order, besides beamforming. We rely on the fact that the MU is already equipped with a temporal equalizer in order to compensate for frequency-selective fading channels, e.g., in macrocells. This equalizer let us take advantage of the channel space diversity by exploiting the spatial decorrelation to emit uncorrelated copies of the desired signal delayed by one or more symbol periods. By doing so, we are transforming the *spatial diversity* at the BS side to *time diversity* at the MU side. This conversion is specially useful when the channel is flat-fading, i.e., the time-spread is zero.

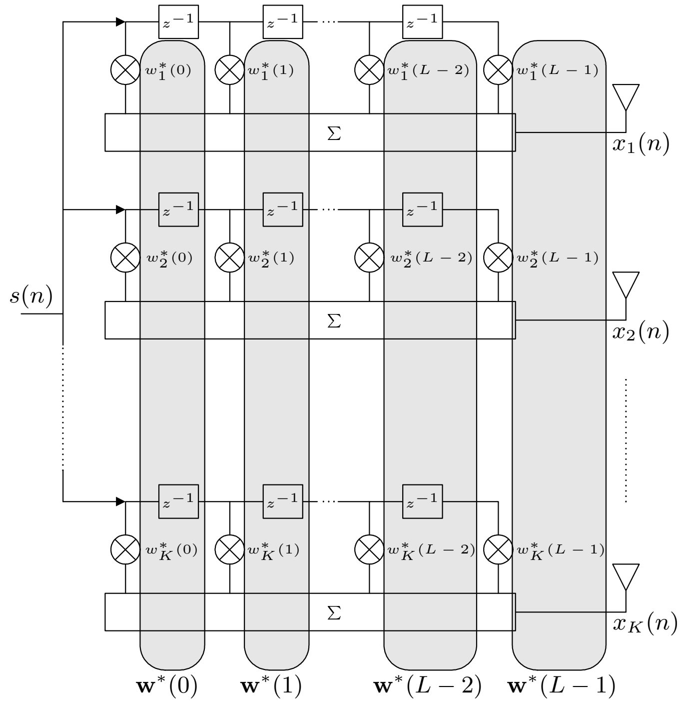

Fig. 1. Transmit Space-Time Filter

The paper is organized as follows. In the next section, the system model used in this work is presented and an expression for the received power by the mobile user is derived. Then, in Section III, we present the proposed criterion and also an adaptive algorithm to find the optimum solution. The results of computer simulations are presented in Section IV, where the performance of the proposed technique is assessed and its behavior is studied. The performance of the proposed algorithm is also compared to more classical techniques. Finally, in Section V, we draw some conclusions and present some open issues.

The following notations are used throughout the paper. Vectors are by default in column orientation, whereas ^{T}, ^{H} and * stand for transpose, conjugate transpose, and conjugate, respectively. $\|\mathbf{x}\|$ is the 2-norm of vector $\mathbf{x}$, $\mathrm{E}\{\cdot\}$ denotes mathematical expectation and $*$ stands for convolution.

II. SYSTEM MODEL

We consider the downlink of a wireless communication system, where the BS is equipped with K antennas and the MU has only one antenna. The TD processing is done by means of a space-time filter, as depicted in Fig. 1. The signal transmitted by the k-th antenna is given by

$$x_k(n) = w_k^*(n) * s(n) = \sum_{l=0}^{L-1} w_k^*(l) s(n-l) \qquad (1)$$

where $s(n)$ are the transmitted symbols to the MU and $w_k(n)$ are the coefficients of the temporal equalizer related to antenna k, which is assumed to have length L.

We assume that the signal is transmitted in blocks of length N_b, so that the channel variation during one block of data is insignificant. However, the channel changes from one block to another, characterizing a *block fading channel*.

At a given block t, the received signal $y(n)$ at the MU antenna can be expressed as

$$y(n) = \sum_{k=1}^{K} h_k(n) * x_k(n) \tag{2}$$

where $h_k(n)$ is the temporal response of the space-time channel, relative to the antenna k, which is assumed to have length D and n is the temporal index within the considered block. We have not represented the index block t for the sake of legibility. This index will be explicitly used in the sequel, when needed.

Rewriting the signal $y(n)$ as

$$y(n) = \sum_{k=1}^{K} \sum_{i=0}^{D-1} h_k(i) x_k(n-i) \tag{3}$$

and recalling that $x_k(n-i) = \sum_{l=0}^{L-1} w_k^*(l) s(n-i-l)$, we obtain that

$$y(n) = \sum_{l=0}^{L-1} \sum_{k=1}^{K} w_k^*(l) \sum_{i=0}^{D-1} h_k(i) s(n-i-l) \ . \tag{4}$$

Defining $\tilde{s}_k(n-l) \triangleq \sum_{i=0}^{D-1} h_k(i) s(n-i-l)$ and the following column vectors (see Fig. 1)

$$\mathbf{w}(l) = \begin{bmatrix} w_1(l) & w_2(l) & \dots & w_K(l) \end{bmatrix}^H \tag{5a}$$

$$\tilde{\mathbf{s}}(n-l) = \begin{bmatrix} \tilde{s}_1(n-l) & \tilde{s}_2(n-l) & \dots & \tilde{s}_K(n-l) \end{bmatrix}^T \tag{5b}$$

we can rewrite (4) as

$$y(n) = \sum_{l=0}^{L-1} \mathbf{w}(l)^H \tilde{\mathbf{s}}(n-l) \ , \tag{6}$$

which can be written in vector notation as

$$y(n) = \mathbf{W}^H \tilde{\mathbf{S}}(n) \tag{7}$$

where

$$\mathbf{W} = \begin{bmatrix} \mathbf{w}(0)^H & \mathbf{w}(1)^H & \dots & \mathbf{w}(L-1)^H \end{bmatrix}^H \tag{8a}$$

$$\tilde{\mathbf{S}}(n) = \begin{bmatrix} \tilde{\mathbf{s}}(n)^T & \tilde{\mathbf{s}}(n-1)^T & \dots & \tilde{\mathbf{s}}(n-L+1)^T \end{bmatrix}^T \ . \tag{8b}$$

The vector $\tilde{\mathbf{S}}(n)$ can be related to the transmitted symbols $s(n)$ in matrix form as shown by (13), where $\mathbf{0}$ is a column vector formed by K zeros and the $K \times D$ matrix $\mathbf{H}_t$ is the space-time response of the channel during the transmission of block t, defined as

$$\mathbf{H}_t = \begin{bmatrix} h_1(0) & h_1(1) & \cdots & h_1(D-1) \\ h_2(0) & h_2(1) & \cdots & h_2(D-1) \\ \vdots & \vdots & \ddots & \vdots \\ h_K(0) & h_K(1) & \cdots & h_K(D-1) \end{bmatrix} \ . \tag{9}$$

Defining the vector of transmitted symbols $\mathbf{s}(n)$ and the block diagonal matrix $\boldsymbol{\mathcal{H}}_t$ as in (13), we can write the received signal at the MU for block t as

$$y_t(n) = \mathbf{W}^H \boldsymbol{\mathcal{H}}_t \mathbf{s}_t(n) \ . \tag{10}$$

It is worth recalling that for each block t the channel presents a different fading condition, i.e., it can be in a deep fade or in a reconstruction condition. This condition affects the signal received power and, at last, the SNR of each received block at the MU.

The average power received by the MU during block t is given by

$$\begin{aligned} P_t &= \frac{1}{N_b} \mathrm{E}\Big\{ y_t(n) y_t^*(n) \Big\} \\ &= \frac{1}{N_b} \mathbf{W}^H \boldsymbol{\mathcal{H}}_t \, \mathrm{E}\Big\{ \mathbf{s}_t(n) \mathbf{s}_t^H(n) \Big\} \boldsymbol{\mathcal{H}}_t^H \mathbf{W} \ . \end{aligned} \tag{11}$$

Assuming that the transmitted symbols $s(n)$ are i.i.d., we obtain $\mathrm{E}\Big\{ \mathbf{s}_t(n) \mathbf{s}_t^H(n) \Big\} = \sigma_s^2 \mathbf{I}$ and (11) becomes

$$P_t = \frac{\sigma_s^2}{N_b} \mathbf{W}^H \boldsymbol{\mathcal{R}}_t \mathbf{W} \ , \tag{12}$$

where $\boldsymbol{\mathcal{R}}_t = \boldsymbol{\mathcal{H}}_t \boldsymbol{\mathcal{H}}_t^H$ is the space-time covariance matrix, which has a block hermitian structure.

III. Constant Power Approach

We assume that the MU has only one antenna and it is already equipped with a temporal equalizer. The main idea is to take advantage of the channel space diversity by exploiting the spatial decorrelation to emit uncorrelated copies of the desired signal (by means of the multiple antenna at the BS) delayed by one or more symbol periods. By doing so, the temporal equalizer at the MU can combine this uncorrelated copies to form an estimation of the transmitted signal. By combining these copies, the MU can profit from the channel diversity to counteract the fading. It is worth noting that

$$\underbrace{\begin{bmatrix} \tilde{\mathbf{s}}(n) \\ \tilde{\mathbf{s}}(n-1) \\ \vdots \\ \tilde{\mathbf{s}}(n-L+1) \end{bmatrix}}_{\tilde{\mathbf{S}}(n)} = \underbrace{\begin{bmatrix} \mathbf{H}_t & & \mathbf{0} & \mathbf{0} & \cdots & \mathbf{0} & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{H}_t & & \mathbf{0} & \cdots & \mathbf{0} & \mathbf{0} & \mathbf{0} \\ \vdots & & & \vdots & & \ddots & & \vdots \\ \mathbf{0} & \mathbf{0} & \mathbf{0} & \mathbf{0} & \mathbf{0} & \cdots & \mathbf{H}_t & \end{bmatrix}}_{\boldsymbol{\mathcal{H}}_t} \underbrace{\begin{bmatrix} s(n) \\ s(n-1) \\ s(n-2) \\ s(n-3) \\ \vdots \\ s(n-D-L+1) \end{bmatrix}}_{\mathbf{s}(n)} \tag{13}$$

the proposed technique transforms the spatial diversity at the BS side into time diversity at the MU side, which can be exploited by a temporal equalizer. Our proposition then takes into account the existent systems and only makes minor changes to increase the diversity gain at the MU side. The performance of the proposed technique is however related to the specific equalizer used at the MU. Here we assume that the MU uses an *ideal* receiver, which is capable of recovering all the signal energy spread across time.

As stated in the previous Section, at each block t a different average power P_t is received by the MU's antenna. The variation of these received powers P_t is caused by the fading. We propose thus to find a (fixed) space-time transmit filter $\mathbf{W}$ that is computed in order that the received power P_t at the MU is as constant as possible over a finite training window. By doing so, we are minimizing the fading effect, i.e., the received power variation. This criterion can be expressed as

$$\mathbf{W}_{\text{opt}} = \arg \min_{\mathbf{W}} \sum_{t=t_i}^{t_f} \left(\mathbf{W}^H \boldsymbol{\mathcal{R}}_t \mathbf{W} - 1\right)^2 \tag{14}$$

where t_i and t_f are the initial and final blocks of the training window, respectively.

In order to keep the same transmit power of the single antenna (no transmit filter) case, the optimum transmit filter is normalized

$$\mathbf{W}_{\text{on}} = \frac{\mathbf{W}_{\text{opt}}}{\|\mathbf{W}_{\text{opt}}\|} \ . \tag{15}$$

This normalized optimum filter $\mathbf{W}_{\text{on}}$ is then used to transmit the subsequent data blocks. In addition, to ensure the desired target SNR at the MU, a power control is done at the BS. We assume that this power control is integrated in the signal $s(n)$, i.e., this signal is already scaled in order that the received power at the MU respects the target SNR.

This novel criterion for transmit diversity is called Constant Power Approach (CPA). This criterion can be optimized by using one of the many optimization techniques existing in the literature. In the following we propose an adaptive algorithm to optimize the proposed criterion. It is not our goal in this work to find the best optimization technique but only to provide one technique in order to assess the performance of the proposed CPA.

A. Obtaining $\mathbf{W}_{opt}$

One can easily associate (14) with a constant modulus (CM) criterion. This criterion has been for long time investigated and some algorithms to optimize it have been proposed in the literature. All these algorithms are based on the following criterion

$$J_{CMA} = \arg \min_{\mathbf{W}} \sum_{n} \left(\mathbf{W}^H \mathbf{x}(n)\mathbf{x}(n)^H \mathbf{W} - 1\right)^2 \ . \tag{16}$$

By comparing (16) and (14), one can easily identify the covariance matrix $\boldsymbol{\mathcal{R}}_t$ with the instantaneous signal covariance matrix $\mathbf{x}(n)\mathbf{x}(n)^H$. Thus, there is a fundamental difference between both criteria since the CMA assumes that the matrix $\mathbf{x}(n)\mathbf{x}(n)^H$ has rank 1 while the matrix $\boldsymbol{\mathcal{R}}_t$ can have (and usually has) rank greater than one.

Then, the already existing algorithms, such as the CMA [15], the ACMA [16] and the finite-interval constant modulus algorithm [17], can not be directly used to optimize (14). It is our belief that these algorithms could be modified to optimize (14). In this work, however, we propose a novel algorithm to preform this task.

Based on (14), let us define the cost function to be minimized as

$$J = \sum_{t=t_i}^{t_f} \left(\mathbf{W}_k^H \mathbf{X}_k(t) - 1\right)^2 \ , \tag{17}$$

where $\mathbf{X}_k(t) = \boldsymbol{\mathcal{R}}_t \mathbf{W}_k$ and k is the iteration index. This criterion can be easily identified as a MSE (mean square error) criterion. The minimization of this criterion can be done iteratively by using the Newton's method [18].

Thus, applying the Newton's method to (17), we obtain that

$$\mathbf{W}_{k+1} = \mathbf{W}_k - \alpha \mathbf{R}_k^{-1} \mathbf{p}_k \tag{18}$$

where the coefficient α is taken less than 1 to avoid the divergence of the algorithm and

$$\mathbf{R}_k = \sum_{t=t_i}^{t_f} \mathbf{X}_k(t)\mathbf{X}_k^H(t) = \sum_{t=t_i}^{t_f} \boldsymbol{\mathcal{R}}_t \mathbf{W}_k \mathbf{W}_k^H \boldsymbol{\mathcal{R}}_t^H \tag{19a}$$

$$\mathbf{p}_k = \sum_{t=t_i}^{t_f} \mathbf{X}_k(t) = \sum_{t=t_i}^{t_f} \boldsymbol{\mathcal{R}}_t \mathbf{W}_k \ . \tag{19b}$$

By taking (experimentally) $\alpha = \frac{1}{4}$, we obtain the proposed Constant Power Algorithm (CPA):

$$\mathbf{W}_{k+1} = \frac{1}{2}\mathbf{W}_k + \frac{1}{2}\mathbf{R}_k^{-1}\mathbf{p}_k \ . \tag{20}$$

As we have said at the beginning of this Section, we just present an algorithm to optimize the criterion of (14). The analytical convergence of this algorithm has not been proved. Simulations results however indicate that the algorithm converges. Indeed, the algorithm has never diverged in all the many cases simulated.

IV. Simulation Results

We consider the downlink of a 120° sector of a cell of a wireless system. The data is transmitted in blocks and we assume, without loss of generality, that the channel realization is independent from one block to another, i.e., the phases of each path varies from one block to another. The BS is equipped with K antennas and, unless specified otherwise, the inter-element distance is $\frac{\lambda_c}{2}$, where λ_c is the carrier wavelength. The transmit space-time filter is $K \times L$ (see Fig. 1), where L is the number of temporal coefficients in each spatial branch. We assume that the instantaneous DCCMs $\boldsymbol{\mathcal{R}}_t$ are perfectly known at the BS, for all t.

In order to assess the performance of the proposed technique, for each *training* block, a different channel was drawn and the corresponding DCCMs $\boldsymbol{\mathcal{R}}_t$ were used as input for the proposed algorithm, which computes the optimum transmit filter $\mathbf{W}_{\text{opt}}$, see (14). The normalized optimum filter $\mathbf{W}_{\text{on}}$, see (15), was then used to compute the received power P_t at the MU for the *data* blocks. With the assumption that the

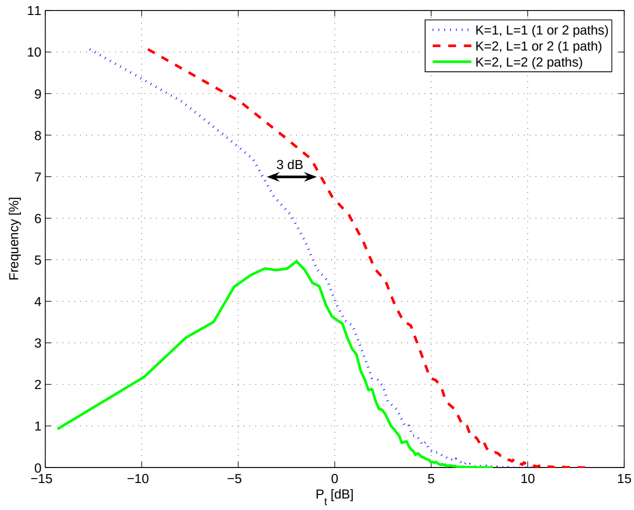

Fig. 2. Histogram of the received power P_t at the MU antenna.

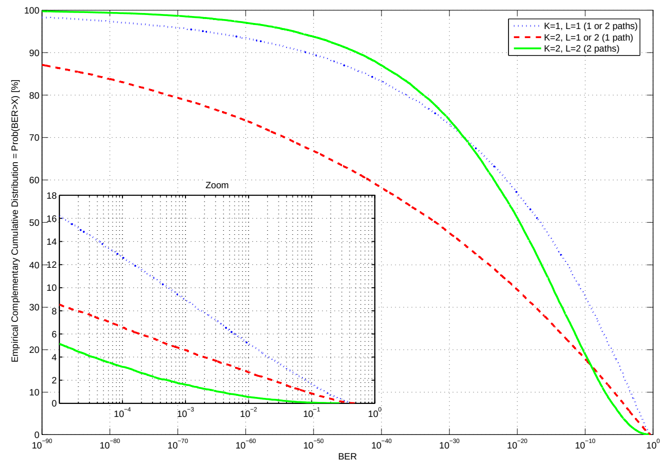

Fig. 3. Outage probability at SNR=20 dB.

temporal equalizer of the MU can recover all the symbol energy, the BER (bit error rate) at its output depends only on the received power P_t and not on the particular equivalent temporal channel. For each block, the theoretical BER for a QPSK modulation was calculated based on the corresponding P_t and the white gaussian noise variance σ^2 at the MU antenna.

We have simulated two different channels, a single path channel and a 2 path channel, in order to highlight the diversity gain provided by the proposed technique. In the 2 paths case, both paths have the same delay but arrives at the receiver from two distinct DOAs (direction of arrival). Moreover, both paths have the same average power of 0.5.

Fig. 2 shows the histogram of the received powers P_t for an SNR of 20 dB. It can be seen that for the 1 path channel, the only improvement between K=1 and K=2 is the 3 dB array gain, i.e., only a shift to the right in the histogram (see the dotted and the dashed curves). In contrast, for the 2 paths channel, the proposed technique with K=2 and L=2 can benefit from the channel's diversity order of 2, which can be seen by the fact that the received power is more concentrated around its (lower) average value.

Fig. 3 shows the outage probability at SNR of 20 dB, which is the probability that the BER is above a given target BER, i.e., no reliable communication is possible. From Fig. 3, one can see that, for the single path case, there is only a scale factor between the K=1 and K=2, which reflects the 3 dB array gain. However, for the 2 paths channel, the curve for K=2 and L=2 (solid curve) shows that the probability of higher BERs is reduced and the BER is more concentrated at lower values, when compared to the single antenna case (dotted curve). The outage probability at 10^{-2} are 5.122 % for K=1, 2.671 % for K=2 (one path) and 0.537 % for K=2 and L=2 (two paths). The proposed technique presents an outage probability which is more than 9 times lower than the single antenna and almost 5 times lower than the two antennas and one path channel case, which means a significant improvement in system capacity.

In order to gain insight into the proposed technique, we have simulated the 2 paths channel, with DOAs of 0° and 90°, for 2 antennas (K=2) and 2 temporal coefficients (L=2).

Fig. 4 shows the radiation pattern of each spatial slice ($\tau = 0$ and $\tau = 1$). It is clear that each pattern points towards one of the channel's path. Due to the delay introduced by the filter, each path will carry a different symbol. Since the symbols of different instants are independent, at the receiver's temporal equalizer, both paths will be summed in power and a diversity gain will be obtained.

In the sequel, we compare the proposed technique (CPA) with the more classical techniques of pure beamforming [19], denoted BF, and the Alamouti scheme [7]. This comparison is carried out in an indoor environment for 3×10^4 blocks. The simulated indoor channel consists of a great number of multipaths with DOA uniformly distributed around the MU antenna. All multipaths have the same propagation delay and thus they arrive at the same instant at the MU antenna. The outage probability of the three techniques (when applicable) at 10^{-2} is showed in Tab. I for different numbers of antennas (K) and temporal coefficients (L). It can be seen that the CPA technique have the same performance as the Alamouti scheme and pure beamforming, but is more flexible than both techniques since the CPA can be applied with any number of antennas and temporal coefficients. Moreover, the CPA technique can be applied regardless of the correlation between the antennas.

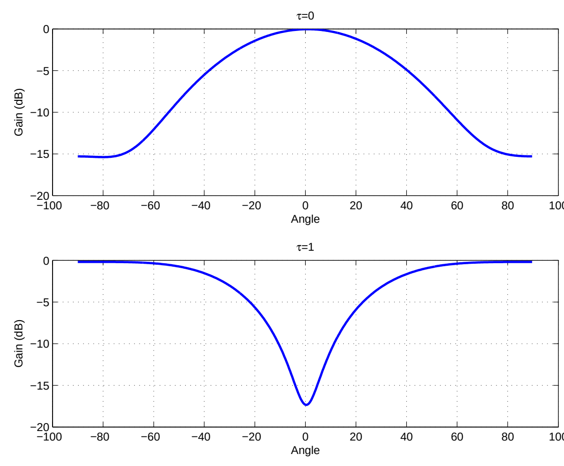

Fig. 4. Radiation patterns for each delay of the transmit filter $\mathbf{W}$.

TABLE I
PERFORMANCE COMPARISON FOR AN INDOOR CHANNEL

		Outage Probability [%] @ BER=10^{-2}		
K	L	CPA	BF	Alamouti
1	1	5.142	5.142	–
1	2	5.142	–	–
2	1	3.999	4.058	–
2	2	0.602	–	0.595

TABLE II
PERFORMANCE RESULTS FOR A FREQUENCY-SELECTIVE FADING CHANNEL

	Outage Probability [%] @ BER=10^{-2}			
	BF	CPA		
	L=1	L=1	L=2	L=3
K=1	2.742	2.742	2.735	2.728
K=2	0.808	0.818	0.815	0.362
K=3	0.665	0.718	0.509	0.109

In order to show the flexibility of the proposed technique, we have simulated the frequency-selective fading channel of [20]. The inter-element distance was set to 1.5λ and the SNR to 14 dB. In this case the Alamouti scheme can not be used due to the temporal spread of the channel. Tab. II shows the outage probability at 10^{-2} for the CPA and BF and for different number of antennas (K) and temporal coefficients (L). It is clear that the CPA with only one temporal coefficient (L=1) is equivalent to pure beamforming and that the inclusion of more temporal coefficients leads to better performances. Significantly better performances are obtained in the case of 3 antennas since the channel model represents a very rich scattering environment, which has a high diversity order. Thus, to profit from this high diversity order, it is necessary to better focus on the individual multipaths. It is worth noting that using the CPA with K=3 and L=3 represents a gain of 6 times in the outage probability with respect to BF with K=3.

V. CONCLUSION

We have proposed a technique called the Constant Power Approach (CPA) that exploits the multiple antennas at the base station to make the received power at the mobile user receiver as constant as possible. This scheme is based on a criterion of the type CM (Constant Modulus). Moreover, we have derived an adaptive algorithm to optimize this criterion. The analytical proof of convergence of this algorithm is however yet an open issue to be investigated. The use of other CM algorithms can also be envisaged.

The simulation results show that the CPA technique presents the same performance as the Alamouti scheme and pure beamforming in equivalent scenarios. The CPA is however more flexible than both techniques since it can be applied with any number of antennas and temporal coefficients. Moreover, the proposed technique can perform jointly beamforming and transmit diversity, which is a very suitable feature for increasing capacity and improving quality in multiuser wireless communication. The Alamouti scheme, on the other hand, creates a higher level of interference by transmitting in an omnidirectional manner.

REFERENCES

[1] W. C. Jakes, *Microwave Mobile Communcations*. New York: IEEE Press, 1974.
[2] M. D. Yacoub, *Foundations of mobile radio engineering*. Boca Raton: CRC Press, 1993.
[3] R. T. Derryberry, S. D. Gray, D. M. Ionescu, G. Mandyam, and B. Raghothaman, "Transmit diversity in 3G CDMA systems," *IEEE Communications Magazine*, vol. 40, no. 4, pp. 68–75, April 2002.
[4] N. Seshadri and J. H. Winters, "Two signaling schemes for improving the error performance of frequency-division-duplex (FDD) transmissions systems using transmitter antenna diversity," in *Proc. IEEE Vehicular Technology Conference*, Secausus, USA, May 1993, pp. 508–511.
[5] ——, "Two Signaling Schemes for Improving the Error Performance of Frequency-Division-Duplex (FDD) Transmission Systems Using Transmitted Antenna Diversity," *International Journal of Wireless Information Networks*, vol. 1, no. 1, pp. 24–47, January 1994.
[6] V. Tarokh, N. Seshadri, and A. R. Calderbank, "Space-time codes for high data rate wireless communications: performance criterion and code construction," *IEEE Transactions on Information Theory*, vol. 44, no. 2, pp. 744–765, March 1998.
[7] S. M. Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE Journal on Select Areas in Communications*, vol. 16, no. 08, pp. 1451–1458, October 1998.
[8] D. Gerlach and A. Paulraj, "Adaptive transmitting antenna arrays with feedback," *IEEE Signal Processing Letters*, vol. 1, no. 10, pp. 150–152, October 1994.
[9] B. Raghothaman, R. T. Derryberry, and G. Mandyam, "Transmit adaptive array without user-specific pilot for 3G CDMA," in *Proc. IEEE International Conference on Acoustics, Speech, and Signal Processing (ICASSP'00)*, vol. 5, Istanbul, Turkey, June 2000, pp. 3009–3012.
[10] T. Asté, P. Forster, L. Féty, and S. Mayrargue, "Downlink Beamforming Avoiding DOA Estimation for Cellular Mobile Communications," in *Proc. IEEE International Conference on Acoustics, Speech, and Signal Processing (ICASSP'98)*, Seattle, USA, May 1998.
[11] T. Asté, L. Féty, P. Forster, and S. Mayrargue, "Downlink Beamforming for Cellular Mobile Communications (GSM system)," *Annales des Télécommunications*, vol. 53, no. 11-12, pp. 435–448, December 1998.
[12] F. Rashid-Farrokhi, K. J. R. Liu, and L. Tassiulas, "Transmit Beamforming and Power Control for Cellular Wireless Systems," *IEEE Journal on Sel. Areas in Communications*, vol. 16, no. 8, pp. 1437–1450, October 1998.
[13] D. Zanatta Filho and L. Féty, "A Fast Algorithm for Joint Downlink Beamforming and Power Control in UMTS Systems," in *Proc. IEEE International Telecommunications Symposium (ITS'2002)*, Natal, Brazil, September 2002.
[14] M. Schubert and H. Boche, "Solution of the Multiuser Downlink Beamforming Problem With Individual SINR Constraints," *IEEE Transactions on Vehicular Technology*, vol. 53, no. 1, pp. 18–28, January 2004.
[15] D. N. Godard, "Self-recovering equalization and carrier tracking in two-dimensional data communication systems," *IEEE Transactions on Communications*, vol. 28, no. 11, pp. 1867–1875, November 1980.
[16] A. van der Veen, *Signal Processing Advances in Wireless and Mobile Communications*. Prentice Hall, 2000, ch. 5 - Algebraic Constant Modulus Algorithms.
[17] P. A. Regalia, "A finite-interval constant modulus algorithm," in *Proc. International Conference on Acoustics Speech, and Signal Processing (ICASSP'2002)*, vol. III, Orlando, USA, May 2002, pp. 2285–2288.
[18] S. Haykin, *Adaptive Filter Theory*, 3rd ed. Prentice Hall, 1996.
[19] L. C. Godara, "Applications of Antenna Arrays to Mobile Communications, Part II: Beam-Forming and Direction-Of-Arrival Considerations," *Proceedings of the IEEE*, vol. 85, no. 8, pp. 1195–1245, August 1997.
[20] P. Pajusco, "Experimental Characterization of D.O.A. at the Base Station in Rural and Urban Area," in *Proc. IEEE Vehicular Technology Conference (VTC'98)*, vol. 2, Ottawa, Canada, May 1998, pp. 993–997.