Power Allocation for OFDM Using Adaptive Beamforming Over Wireless Networks

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***Abstract*—The performance of a multiuser wireless network using orthogonal frequency-division modulation (OFDM), combined with power control and adaptive beamforming for uplink transmission is presented here. A network-wide adaptive power control algorithm is used to achieve the desired signal-to-interference-and-noise-ratio at each OFDM subcarrier and increase the power efficiency of the network. As a result, we can achieve a better overall error probability for a fixed total transmit power. With the assumption of fixed-modulation for all subcarriers, transmit powers and beamforming weight vectors at each subcarrier are updated jointly, using an iterative algorithm that converges to the optimal solution for the entire network. Unlike most of the loading algorithms, this approach considers fixed bit allocation and optimizes the power allocation and reduces the interference for the entire network, rather than a single transmitter. We also propose joint time-domain beamforming and power control to reduce the complexity resulting from the number of beamformers and fast Fourier tranformed blocks. The proposed algorithm is also extended to COFDM and we show that it improves the performance of those systems.**

***Index Terms*—Antenna array, coded orthogonal frequency-division modulation (COFDM), minimum mean-squared error (MMSE), minimum variance distortionless response (MVDR), OFDM, power control.**

I. Introduction

ORTHOGONAL frequency-division multiplexing (OFDM) is a parallel data transmission scheme. If the width of each subcarrier is smaller than the coherence bandwidth of channel, it converts the wideband frequency-selective fading channel to a series of narrowband flat-fading subchannels [1]. For a small tone spacing, relative to the coherence bandwidth of channel, there is no need for sophisticated equalization methods [2]. A disadvantage of OFDM is that overall bit-error rate (BER) is dominated by the performance of the worst signal-to-noise ratio (SNR) subcarrier [3], [4]. If the signal-to-interference-and-noise-ratio (SINR) fluctuates over subcarriers, the ones with the worst SINR would affect the overall BER the most. As a result, in the case of frequency selective fading channels, the error probability of the whole system will improve slowly by increasing the transmitted power. In order to obtain a minimum overall error probability, the optimum algorithm is to have a uniform error probability for all of the subcarriers [3].

Several schemes have been proposed to combat the aforementioned problem. One solution is to use coded OFDM (COFDM) [5]–[7]. Other methods try to adjust the bit and power distribution among subcarriers according to their link gains and are mostly called "loading algorithms" [4], [8]–[10]. In most of these algorithms, the bit allocation in each subcarrier is adapted to its capacity and therefore, a fixed modulation scheme is not considered for all subcarriers. Hughes-Hartogs [10] proposed a method in which the bits are assigned one by one to the subcarrier with the lowest power increment, until a prespecified target rate R_T is reached. Obviously this is a very slow procedure and requires lots of sorting and searching. Chow and Cioffi [8] distribute the bits according to the capacity of the subcarriers, by using the concept of "SNR gap approximation." Another loading algorithm has been proposed in [4] where the distribution of bits is adapted to the shape of the transfer function of each subcarrier. Fischer and Huber [9] exploit the fact that the signal power and the rate at each subcarrier are related. They minimize the BER at each subcarrier with a constant data rate and transmission power.

However, most of these methods have been proposed for a single user system without considering the effect of interferences. As a result, they cannot reach the optimum solution in the sense of minimum total transmission power. Moreover, the loading in one transmitter can change the interference in other receivers, and therefore an iterative procedure is needed. In contrast, we use an adaptive power allocation scheme to distribute transmit power among subcarriers based on the interference from other users at the same subcarriers, assuming fixed modulation for all of them (used in systems like IEEE802.11 or IEEE802.16). In a fast varying channel, implementing bit-loading algorithms requires a large bandwidth to feedback the varying SNR at each subcarrier to the transmitter. Here, we assume the modulation for all subcarriers to be fixed, and by adjusting the transmit power at each subcarrier through a slow feedback link, we try to compensate for slow fading. Furthermore, we exploit antenna arrays to perform both frequency and time-domain beamforming to further reduce the interference. If the array response is not known at the receiver, we use minimum mean-squared error (MMSE) beamforming, in which training sequences are used to update the weight vectors and minimize the interference. The rate at which the training sequences are transmitted depends on the speed of channel variation. It should be noted here that sometimes because of deep frequency selective fading, some

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tones experience very low SNR, and a tremendous amount of power adjustment might be needed to achieve the desired SNR. In this case, it might be better to discard the subcarrier, instead of allocating a large portion of power. However, this can result in data rate reduction. Another approach is to truncate the power of each subcarrier at a prespecified level.

In COFDM, by coding across subcarriers, the effect of fading is averaged over all of them. However, by exploiting the power control and beamforming, the SINR at all of the subcarriers can be increased, and therefore, the overall BER is decreased. Moreover, power control and beamforming can reduce the total network power. We will compare the performance of an uncoded OFDM using our proposed algorithms with that of COFDM systems with and without power control or beamforming.

In this paper, we will assume that each mobile uses all of the subcarriers. However, with a slight modification, the same formulations and the same algorithms can be applied to OFDMA, where the subcarriers are partitioned and each partition is assigned to a (group of) user(s).

This paper is organized as follows. In Section II, we will review the concept of power control and propose the power control for OFDM receivers. Section III proposes the OFDM joint power control and frequency-domain beamforming. Joint time-domain beamforming and power control is proposed in Section IV. In Section V, we will use MMSE approach to perform the beamforming. Section VI extends the proposed algorithm to COFDM. Section VII presents some simulation results, and finally Section VIII concludes the paper.

II. System Model

The objective of power control in wireless networks is to minimize the transmitted power while some target error probabilities are met [11], [12]. Consider a network of M mobiles trying to access the same channel. We denote the power link gain between the ith mobile and the bth base station by a real number G_{ib}, and the ith mobile transmitted power by P_i. we assume that one base station is assigned to each mobile. Moreover, these base stations use the same frequency band and so they suffer from co-channel interference. The SINR at the bth receiver is given by

$$\Gamma_b = \frac{G_{bb}P_b}{\sum\limits_{\substack{i=0\\i\neq b}}^{M-1} G_{ib}P_i + N_b} \tag{1}$$

where N_b is the noise power at the bth base station. The objective is to maintain the total transmitted power as low as possible, while the SINRs are kept above a threshold. If we denote the minimum acceptable SINR at base b by γ_b, the lowest possible total power is obtained when all of the SINRs are equal to the threshold, i.e. $\Gamma_b = \gamma_b, b = 0, \ldots, M-1$.

A distributed power update scheme is proposed in [13] that achieves the optimal solution for (2). The bth mobile power at the nth stage of iteration is updated by

$$P_b(n+1) = \frac{\gamma_b}{G_{bb}} \left(\sum_{\substack{i=0\\i\neq b}}^{M-1} G_{ib}P_i(n) + N_b \right). \quad b = 0, \ldots, M-1 \tag{2}$$

The right-hand side in (2) is a function of the noise and interference at the bth base station (the term inside parenthesis), the link gain G_{bb}, and the target SINR. All of these can be measured locally and transmitted through a feedback channel to the corresponding mobile [11].

In the following, we consider this scheme in a multiuser environment using multicarrier transmission. Our objective is to optimize the power allocation at all fixed-modulated subcarriers for all of the mobiles, such that: *a*) The SINR at all of the subcarriers for all of the mobiles are close to each other and they are above a SINR threshold. *b*) The total power used to achieve the aforementioned objective is minimized. The basic idea is to allocate less power to the subcarriers with less interference, and more power to the subcarriers with lower SINR. If the maximum number of paths between the ith user and the bth base station is assumed to be L, the corresponding link can be modeled by the following impulse response (we have ignored the Doppler effect):

$$h_{ib}(t) = \sqrt{G_{ib}} \sum_{l=0}^{L-1} \alpha_{ib}^l \delta\left(t - \tau_{ib}^l\right) \tag{3}$$

where α_{ib}^l denotes the lth path fadings that are independent complex Gaussian variables with variance ${\sigma_{ib}^l}^2$ (their amplitudes are Rayleigh); τ_{ib}^l's are the delays of the corresponding paths; and G_{ib} is a real random variable representing the log-normal shadow fading and path loss.

In this paper, the vectors are shown by bold underline letters. Moreover, the transmitted and received signals at the time domain are shown by uppercase and the same values at the frequency domain by lowercase letters. Let us assume that N denotes the number of subcarriers, T_s the symbol period, and f_0 the carrier frequency. If N is large enough, each subcarrier can be modeled as a flat fading channel [1], and so the link gain at subcarrier c, H_{ib}^c can be calculated simply by replacing f with $f_c \triangleq f_0 + (c/NT_s)$ in $H(f)$, the Fourier transform of $h_{ib}(t)$ in (2), i.e.,

$$H_{ib}^c = \sqrt{G_{ib}} \sum_{l=0}^{L-1} \alpha_{ib}^l e^{\left(-j2\pi f_c \tau_{ib}^l\right)}. \tag{4}$$

Without loss of generality, we can assume the path loss and shadowing for different paths to be the same, and any difference can be absorbed in fading coefficients.

In this paper, we assume that a proper guard interval has been inserted in time domain such that the effect of intersymbol interference (ISI) can be ignored. Moreover, the guard interval has the form of cyclic prefix and, therefore, the interaction of the received signals at different subcarriers in the frequency domain is zero. This is due to the cyclic convolution performed between the channel and the transmitted signal. The modulated data at subcarrier c for user i is d_i^c whose energy is assumed to be unity (e.g. using a fixed MPSK on all subcarriers).

We assume that all of the subcarriers use the same modulation and so due to the fact that subcarrier link gains are different, by distributing the power equally among them, the SINRs would become unbalanced. Now, we perform a power control algorithm at each subcarrier separately. If we denote the power al-

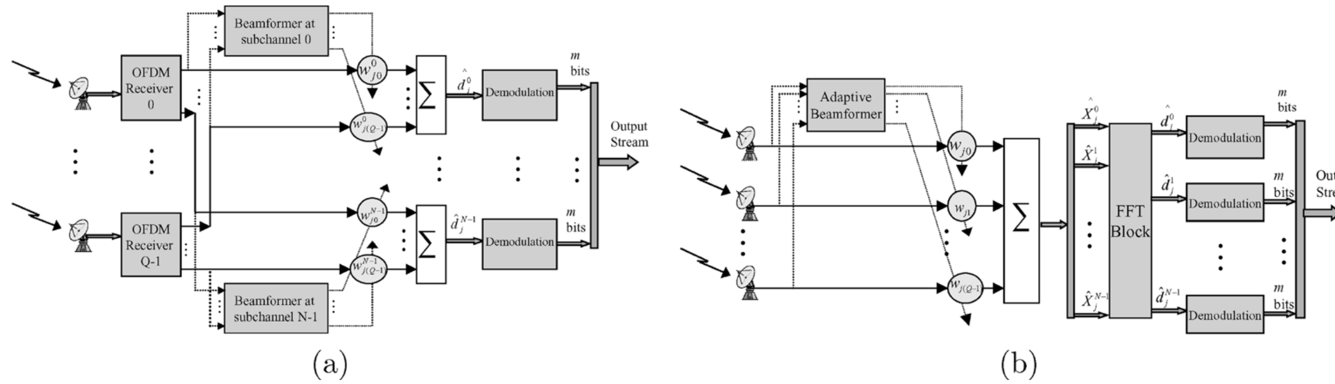

Fig. 1. (a) Frequency-domain and (b) time-domain beamforming in the jth OFDM receiver.

located to mobile i at subcarrier c by P_i^c and define $W_N \triangleq e^{-j2\pi/N}$, the kth sampled received signal $(k = 0 \ldots N-1)$ at the bth receiver (after down conversion, guard interval removal, proper matched filtering, and sampling at intervals T_s) will be

$$\begin{aligned}\hat{X}_b^k &= \frac{1}{\sqrt{N}} \sum_{i=0}^{M-1} \left(\sum_{c=0}^{N-1} \underbrace{H_{ib}^c \sqrt{P_i^c} d_i^c}_{t_{ib}^c} W_N^{-ck} \right) + n_b^k \\ &= \frac{1}{\sqrt{N}} \sum_{c=0}^{N-1} \left(\sum_{i=0}^{M-1} t_{ib}^c + \hat{n}_b^c \right) W_N^{-ck} \end{aligned} \tag{5}$$

where n_b^k is the kth noise sample and $\hat{n}_b^c$ is the Fourier transform of the noise samples. Since the noise samples are uncorrelated, these two variables have the same power, N_b.

The signal at subcarrier c for base station b in frequency domain is

$$\hat{d}_b^c = \sum_{i=0}^{M-1} t_{ib}^c + \hat{n}_b^c = t_{bb}^c + \left[\sum_{i \neq b}^{M-1} t_{ib}^c + \hat{n}_b^c \right], \; i = 0, \ldots, N-1 \tag{6}$$

where the first part is the desired signal attenuated by the link gain, and the term inside the bracket is the sum of the interferences and thermal noise. The SINR at the cth subcarrier Γ_b^c is given by (2) as a function of link gain, power value, and noise at the cth subcarrier.

Our goal is to maintain Γ_b^c above a target value γ_b^c while the sum of allocated powers is minimized. To achieve this goal, we apply the power control algorithm, described in the previous subsection, to each subcarrier independently. Since the subcarriers are assumed to be orthogonal, this guarantees that the SINR at each subcarrier is at least γ_b^c [11], [12].

III. Power Control and Frequency-Domain Beamforming

Now consider an uplink OFDM system where adaptive beamforming is deployed at each subcarrier of all OFDM receivers. Fig. 1(a) depicts the receiver with frequency-domain beamforming at each subcarrier which ensures that the subcarriers can still be considered independently. The kth sampled received vector at the bth base station at the time domain is given by (using the notations introduced in (2))

$$\underline{\mathbf{X}}_b^k = \frac{1}{\sqrt{N}} \sum_{i=0}^{M-1} \left(\sum_{c=0}^{N-1} t_{ib}^c W_N^{-ck} \right) \underline{\mathbf{a}}_{ib} + \underline{\mathbf{n}}_b^k \tag{7}$$

where the Q-element vector $\underline{\mathbf{a}}_{ib}$ is the array response at the bth receiver for the ith transmitter, and $\underline{\mathbf{n}}_b^k$ is the noise vector (with dimension Q, the number of antennas) whose elements are the noise samples at the input of each antenna.

The resultant cth outputs of the FFT blocks create the vector $\hat{\underline{\mathbf{d}}}_b^c = \sum_{i=0}^{M-1} t_{ib}^c \underline{\mathbf{a}}_{ib} + \hat{\underline{\mathbf{n}}}_b^c$, where $\hat{\underline{\mathbf{n}}}_b^c$ is the Fourier transform of $\underline{\mathbf{n}}_b^k$. The output of the beamformer at subcarrier c is then given by $e^c = \underline{\mathbf{w}}_b^{cH} \hat{\underline{\mathbf{d}}}_b^c$.

If we assume that the receiver knows the array response to the desired user, we can use minimum variance distortionless response (MVDR) beamforming [14]. In MVDR, the weight vector is calculated in order to minimize the total energy at the beamformer output, when the gain toward the desired direction is fixed. The joint beamforming and power control algorithm is performed at each subcarrier separately, assuming perfect orthogonalization. The energy of the beampattern at subcarrier c is $\epsilon^c = E[e^c e^{c*}] = \underline{\mathbf{w}}_b^{cH} E[\hat{\underline{\mathbf{d}}}_b^c \hat{\underline{\mathbf{d}}}_b^{cH}] \underline{\mathbf{w}}_b^c$. Assuming that the noise is zero mean, white Gaussian process, and the transmitted symbols are independent and have average unity energy (see the assumptions in Section II), we obtain

$$\epsilon^c = \underbrace{\left[\sum_{i \neq b} \left(P_i^c \left| H_{ib}^c \right|^2 \left| \underline{\mathbf{w}}_b^{cH} \underline{\mathbf{a}}_{ib} \right|^2 \right) + N_b \left| \underline{\mathbf{w}}_b^c \right|^2 \right]}_{I_b^c} + P_b^c \left| H_{bb}^p \right|^2 \left| \underline{\mathbf{w}}_b^{cH} \underline{\mathbf{a}}_{bb} \right|^2 \tag{8}$$

where I_b^c is the interference plus noise and the second term is the power of the signal coming from the desired direction.

The SINR at the output of the beamformer at subcarrier c is given by

$$\Gamma_b^c = \frac{P_b^c |H_{bb}^c|^2 \left| \underline{\mathbf{w}}_b^{cH} \underline{\mathbf{a}}_{bb} \right|^2}{\sum\limits_{i \neq b} \left(P_i^c |H_{ib}^c|^2 \left| \underline{\mathbf{w}}_b^{cH} \underline{\mathbf{a}}_{ib} \right|^2 \right) + N_b \left| \underline{\mathbf{w}}_b^c \right|^2}. \quad c = 0, \ldots, N-1. \tag{9}$$

The MVDR solution for beamforming optimization will be [14]

$$\underline{\mathbf{w}}_b^c = \frac{(\mathbf{R}_b^c)^{-1}\underline{\mathbf{a}}_{bb}}{\underline{\mathbf{a}}_{bb}^H(\mathbf{R}_b^c)^{-1}\underline{\mathbf{a}}_{bb}}, \quad c = 0, \dots, N-1 \tag{10}$$

where the data correlation matrix at the cth subcarrier of base station b is

$$\mathbf{R}_b^c = \sum_{i \neq b} \left(P_i^c |H_{ib}^c|^2 \underline{\mathbf{a}}_{ib}\underline{\mathbf{a}}_{ib}^H \right) + N_b \mathbf{I}. \tag{11}$$

Note that in (3) we have used $i \neq b$. Using the Matrix Inversion Lemma [14], it is straightforward to show that by choosing $\underline{\mathbf{w}}_b^c$ as in (3), we can drop the restriction $i \neq b$ in the definition of autocorrelation matrix in MVDR beamforming.

By considering the fact that the MVDR constraint enforces $|\underline{\mathbf{w}}_b^{cH}\underline{\mathbf{a}}_{bb}|^2 = 1$, we solve (3) in terms of P_b^c and adopt the iterative scheme presented in [11]. As a result, the mapping at each iteration is the combination of (3) and the following equation:

$$P_b^c(n+1) = \frac{\gamma_b^c}{|H_{bb}^c|^2} I_b^c(n). \tag{12}$$

The following algorithm achieves the jointly optimal power allocations and adaptive beamforming, assuming the same modulation at all subcarriers.

1) At step $n = 0$, the bth base station sets $P_b^c(n) = 0$ ($c = 0 \dots N-1$) for its mobile.
2) For each subcarrier, the bth base station calculates the autocorrelation matrix $\mathbf{R}_b^c$, and uses (3) to compute the weight vector $\underline{\mathbf{w}}_b^c$.
3) The base station calculates the interference and noise at subcarrier c, $I_b^c(n)$, as given in (3), and transmits it to the transmitter through the feedback channel.
4) The mobile transmitter updates the power at each subcarrier according to (3).
5) If $P_b^c(n+1) > P_{\max}$, we set $P_b^c(n+1) = P_{\max}$, where $P_{\max}$ is a predetermined maximum power. This prevents the subcarriers in deep fade to consume a tremendous amount of power
6) If $\sum_{p=0}^{N-1} |P_b^p(n+1) - P_b^p(n)|^2 \leq \mu$, when μ is a threshold that defines the speed of convergence, the base station stops, otherwise sets $n = n+1$ and goes back to step 2.

Note that instead of imposing an upper bound on each subcarrier's power ($P_{\max}$), we can discard the subcarrier which is in deep fade. However, this results in data rate reduction.

If there is a solution for the joint power control and beamforming problem, this algorithm will converge to the optimum solution and this solution is unique [11] (assuming a fixed modulation).

In this work, we assumed that each mobile is using all available subcarriers. However, in orthogonal frequency-division multiple access (OFDMA), users are grouped and a subset of OFDM subcarriers is assigned to each (group of) user(s). In this case, the proposed algorithms can be applied to the co-channel users in each group by replacing N with the number of subcarriers assigned to a group. The modulation schemes can be different in different groups and this allows us to have a separate desired SINR for each group.

IV. Time-Domain Beamforming

The complexity (number of multiplications) of frequency-domain beamforming at the receiver is in the order of $QN\log N + NQ^4$. In the system depicted in Fig. 1(b), the beamforming is performed in the time domain and unlike Fig. 1(a), only one set of weight vectors is calculated at each iteration rather than N sets. The complexity of this system is $N\log N + Q^4$, which is significantly less compared with that of Fig. 1(a). With $Q = 4$ and $N = 128$, the complexity of Fig. 1(a) is in the range of 33847, while that of Fig. 1(b) is 525, a complexity decrease of an order of 64.

Using the system depicted in Fig. 1(b), we are no longer able to consider the joint beamforming and power control at each subcarrier independently. In an OFDM system, the symbol decisions are made at the FFT output. The error and weight vector calculations have to be done in the frequency domain. If a time-domain beamformer is to be used, we need to relate the frequency domain error to that quantity in the time domain. One way to look at this problem is to minimize the energy of $\hat{D}_b$, the output of the beamformer in Fig. 1(b). Using the Parseval equation, this is equivalent to minimizing the sum of the energies of the subcarriers at the output of the FFT block. If we denote the kth sample input to the FFT block at receiver b by $\hat{X}_b^k$, then using (3), the received signal at the pth subcarrier will be

$$\hat{d}_b^p = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} \hat{X}_b^k W_N^{kp} = \sum_{i=0}^{M-1} t_{ib}^p \sum_{q=0}^{Q-1} a_{ib}^q w_{bq}^* + \sum_{q=0}^{Q-1} w_{bq}^* \hat{n}_{bq}^p \tag{13}$$

where t_{ib}^p is defined as in (2). Note that in (4) we have used the property of Comb Sequences, which states $\sum_{k=0}^{N-1} W_N^{kp} = 0$ for all $p \neq 0$. Using the fact that the input symbol energy is unity, the signal energy at subcarrier p is obtained by

$$e^p = E\left[\hat{d}^p \hat{d}^{p^*}\right] = \sum_{i=0}^{M-1} P_i^p |H_{ib}^p|^2 \left|\underline{\mathbf{w}}_b^H \underline{\mathbf{a}}_{ib}\right|^2 + \frac{N_b}{2} |\underline{\mathbf{w}}_b|^2.$$

It is not possible to minimize the energy of all of the subcarriers simultaneously, thus we use a metric which is a positive combination of all e^p's. Since each e^p is actually an energy quantity, we simply minimize the sum of the energies that is equivalent to the energy at the output of the beamformer $\hat{D}_b$, as illustrated in Fig. 1(b). So our optimization problem becomes

$$\underline{\mathbf{w}}_b = \arg\min_{\underline{\mathbf{w}}_b} \left\{ \sum_{i=0}^{M-1} \left|\underline{\mathbf{w}}_b^H \underline{\mathbf{a}}_{ib}\right|^2 \sum_{p=0}^{N-1} P_i^p |H_{ib}^p|^2 + N N_b |\underline{\mathbf{w}}_b|^2 \right\},$$
$$\text{subject to } \underline{\mathbf{w}}_b^H \underline{\mathbf{a}}_{bb} = 1$$

where the term inside the bracket is equal to $\sum_{p=0}^{N-1} e^p$.

This is very similar to a normal beamforming process. The solution for the vector $\underline{\mathbf{w}}_b$ is similar to (3) with $\mathbf{R}_b$ redefined as

$$\mathbf{R}_b = \sum_{i \neq b} \left(\underline{\mathbf{a}}_{ib}\underline{\mathbf{a}}_{ib}^H \sum_{p=0}^{N-1} P_i^p |H_{ib}^p|^2 \right) + N N_b \mathbf{I} = \sum_{p=0}^{N-1} \mathbf{R}_b^p \tag{14}$$

where $\mathbf{R}_b^p$ is the autocorrelation matrix at subcarrier p as defined in (3). As in the case of frequency-domain beamforming, the restriction $i \neq b$ in (4) can be dropped.

By replacing $\underline{\mathbf{w}}_b^c$ with $\underline{\mathbf{w}}_b$, the SINR at the cth subcarrier can be evaluated by (3), and at each receiver the iterative algorithm would be the same as the one for frequency-domain beamforming presented in Section III, except in step 2, where the bth base station calculates the sum of autocorrelations of all subcarriers, $\sum_{c=0}^{N-1} \mathbf{R}_b^c$. Moreover, instead of $\underline{\mathbf{w}}_b^c$, we calculate and use $\underline{\mathbf{w}}_b$.

Let us assume that the gain matrix is denoted by $\mathbf{F}(\underline{\mathbf{w}})$ whose (ib)th element is $(\gamma_b^c |H_{ib}^c|^2 |\underline{\mathbf{w}}_b^H \underline{\mathbf{a}}_{ib}|^2)/(|H_{bb}^c|^2)$ for $i \neq b$ and 0 for $i = b$. The gain matrix is an irreducible nonnegative definite matrix, and by the Perron–Frobenius theorem [15], has a positive real eigenvalue that is larger than the amplitude of all other eigenvalues (spectral radius of the matrix). If the spectral radius of the gain matrix is less than unity, there is a solution for the algorithm [11], [12]. Let us call the mappings defined by the modified version of (3) (replacing $\underline{\mathbf{w}}_b^c$ with $\underline{\mathbf{w}}_b$) $m^w(\mathbf{p}^n)$, and the mapping defined by the combination of (4) and (3) as $m(\mathbf{p}^n)$. Since the coefficients of the power values in these mappings are positive, [11, Th. 1] is applicable to prove the convergence and optimality of this algorithm. The fixed point of the mapping $m(.)$ is unique [11]. Therefore, if the link gains and steering vectors are such that there exists a solution for this joint power control and beamforming problem, the above mentioned algorithm will always converge to a unique optimal solution. If there is a solution to the iterative algorithm, the application of the upper bound to each subcarrier's power ($P_{\max}$) will expedite the convergence.

Like frequency domain, in time-domain beamforming, only one real value is exchanged through the feedback channel from the receiver to the transmitter for each link per update. Therefore, the required bandwidth for the feedback channel is the same for both methods.

V. MMSE Beamforming Approach

If the base stations do not have full knowledge of the array responses, $\underline{\mathbf{a}}_{ib}$, we must use a training sequence which is correlated with the desired signal. The weight vector is obtained by minimizing the MSE between the estimated signal and the training sequence. This is called MMSE approach [16]. MMSE can be applied to both frequency and time-domain beamforming. Here, we only show the MMSE time-domain beamforming. If we call the kth sample at the qth antenna at base station b by X_{bq}^k, from (4) we obtain

$$\begin{aligned} \hat{d}_b^c &= \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} \hat{X}_b^k W_N^{kc} \\ &= \sum_{q=0}^{Q-1} w_{bq}^* \left[\frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_{bq}^k W_N^{kc} \right] \\ &= \underline{\mathbf{w}}_b^H \underline{\mathbf{s}}_b^c \end{aligned} \tag{15}$$

where the sequence s_{bq}^c $(c = 0 \ldots N-1)$ is the Fourier transform of the sequence X_{bq}^k $(k = 0 \ldots N-1)$. The objective in MMSE beamforming is to minimize ε_{cb}^2 given by

$$\varepsilon_{cb}^2 = E\left[\left|\hat{d}_b^c - t_b^c\right|^2\right] = P_b^c |H_{bb}^c|^2 + \underline{\mathbf{w}}_b^H \mathbf{R}_{ss}^p \underline{\mathbf{w}}_b - 2\mathrm{Re}\{\underline{\mathbf{w}}_b^H \underline{\mathbf{r}}_{st}^c\} \tag{16}$$

where $t_b^c = \sqrt{P_b^c} H_{bb}^c d_b^c$, and d_b^c is the training sequence at the bth transmitter whose power is assumed to be unity. Moreover, $\mathbf{R}_{ss}^c = E[\underline{\mathbf{s}}_b^c \underline{\mathbf{s}}_b^{cH}]$ and $\underline{\mathbf{r}}_{st}^c = E[\underline{\mathbf{s}}_b^c t_b^{c*}]$.

Here we have N subcarriers and the weight vector is the same for all of them, so the criteria in MMSE time-domain beamforming is to minimize $\sum_{c=0}^{N-1} \varepsilon_{cb}^2$. This is a typical MMSE optimization problem and if $\sum_{c=0}^{N-1} \mathbf{R}_{ss}^c$ is nonsingular, its solution will be the well-known Wiener–Hopf equation [16] given by

$$\underline{\mathbf{w}}_b^{\mathrm{opt}} = \left(\sum_{c=0}^{N-1} \mathbf{R}_{ss}^c \right)^{-1} \sum_{c=0}^{N-1} \underline{\mathbf{r}}_{st}^c. \tag{17}$$

The time-domain beamformer does not minimize the individual errors at each subcarrier. Therefore, the principal of orthogonality, valid for the Wiener-Hopf solution, is not satisfied here. However, in the following lemma, we will prove that this solution is equivalent to the MVDR solution up to a constant coefficient and, therefore, results in the same SINR [16].

Lemma: If the training sequences transmitted from different mobiles are uncorrelated, the MMSE weight vector presented in (5) is equivalent to the MVDR weight vector, as expressed in (3) with autocorrelation matrix defined as in (4), up to a constant coefficient.

Proof: Using (3) and the definition of vector $\underline{\mathbf{s}}_b^c$, we have

$$\begin{aligned} \underline{\mathbf{r}}_{st}^c &= E\left[\underline{\mathbf{s}}_b^c t_b^{c*}\right] \\ &= \frac{1}{N} \sum_{i=0}^{M-1} \sum_{p=0}^{N-1} \underline{\mathbf{a}}_{ib} E\left[t_i^p t_b^{c*}\right] \sum_{k=0}^{N-1} W_N^{k(c-p)} \\ &\quad + \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} E\left[\underline{\mathbf{n}}_b t_b^{c*}\right] W_N^{kc} \\ &= \frac{1}{N} \sum_{i=0}^{M-1} \underline{\mathbf{a}}_{ib} E\left[t_i^c t_b^{c*}\right] \\ &= E\left[\left|t_b^c\right|^2\right] \underline{\mathbf{a}}_{bb} \\ &= P_b^c \left|H_{bb}^c\right|^2 \underline{\mathbf{a}}_{bb}. \end{aligned}$$

On the other hand, by using the definition of $\mathbf{R}_{ss}^c$, we get

$$\begin{aligned} \mathbf{R}_{ss}^c &= E\left[\underline{\mathbf{s}}_b^c \underline{\mathbf{s}}_b^{cH}\right] \\ &= \sum_{i=0}^{M-1} \sum_{i'=0}^{M-1} \underline{\mathbf{a}}_{ib} \underline{\mathbf{a}}_{i'b}^H E\left[t_i^c t_{i'}^{c*}\right] + \frac{1}{N} \sum_{k=0}^{N-1} E\left[\underline{\mathbf{n}}_b^k \underline{\mathbf{n}}_b^{k^H}\right] \\ &= \sum_{i=0}^{M-1} \underline{\mathbf{a}}_{ib} \underline{\mathbf{a}}_{ib}^H P_i^c \left|H_{ib}^c\right|^2 + N_0 \mathbf{I} = \mathbf{R}_b^c. \end{aligned}$$

This is the same as the autocorrelation for MVDR defined in (4). Using (5) and (5), the weight vectors for MVDR and MMSE time-domain beamforming are the same. ■

The interference at subcarrier c equals the difference between the received power, $E[|\hat{d}_b^c|^2]$, and the power of the desired signal and is given by

$$I_b^c = \underline{\mathbf{w}}_b^{\mathrm{opt}\,H} \left(\mathbf{R}_{ss}^c - \frac{1}{P_b^c \left|H_{bb}^c\right|^2} \underline{\mathbf{r}}_{st}^c \underline{\mathbf{r}}_{st}^{c\;H} \right) \underline{\mathbf{w}}_b^{\mathrm{opt}}. \tag{18}$$

Therefore, the MMSE algorithm is outlined as follows.

1) At step $n = 0$, the bth base station sets $P_b^c(n) = 0$; $(c = 0 \dots N-1)$ for its mobile.
2) Using (5), the bth base station finds the weight vector $\underline{\mathbf{w}}_b^{\mathrm{opt}}$.
3) The base station uses (5) to find the interference at each subcarrier and transmits these values to the bth mobile through the feedback channel.
4) The bth mobile uses (3) to re-calculate the power at each subcarrier. If $P_b^c(n+1) > P_{\max}$, we set $P_b^c(n+1) = P_{\max}$.
5) The base station sets $n = n+1$ and repeats from step 2 until convergence.

VI. Extension to COFDM

Theoretically, the bit and power allocation obtained by the loading algorithms meet the desired BER as long as the time variation of the channel is very limited. Performing bit loading on time-varying channels requires a mechanism to adapt to the channel variation. Many practical OFDM systems use coding across subcarriers (in frequency domain) to achieve better immunity to the frequency-selective fading channels. This provides a link between bits transmitted on separate subcarriers and is done in such a way that the information conveyed by the subcarriers in deep fade can be reconstructed by the information received through the ones with good channels. Block or convolutional codes are used either by their own or combined together (as the inner and outer code) and possibly with interleaving. In trellis-coded modulation (TCM), convolutional coding is combined with modulation and results in higher coding gain. Mostly, TCM is based on the set partitioning performed by the Ungerboeck's encoder [17], in which m information bits map into a signal from the 2^{m+1}-ary constellation. k of these bits are encoded by a rate-$k/(k+1)$ convolutional encoder to select one of the 2^{k+1} partitions at the $(k+1)$th level of the constellation's partition tree. The remaining $m-k$ bits are used to select one point within the designated partition. Adaptive TCM (ATCM) uses MQAM constellation and has a coding gain of at least 7 dB over simple TCM (see [6] for details).

Although the COFDM averages the fading over all subcarriers, in an environment with low or moderate high SNR, the BER depends on the SNR of each subcarriers. This dependence could be better seen for TCM, through the following inequality [18], [19]:

$$N_{d_f} Q\left(\sqrt{\frac{d_f^2 E_s}{(2N_0)}}\right) \le P_e \le \sum_{d=d_f}^{\infty} N_d Q\left(\sqrt{\frac{d^2 E_s}{(2N_0)}}\right) \quad (19)$$

where d_f^2 is the normalized square free distance of the code, N_d and N_{d_f} are the average number of paths in the trellis having the squared Euclidian distance of d^2 and d_f^2 from the all-zero path, respectively, and $Q()$ is the error function.

Moreover, in a multiuser environment it is not only the fading that determines both the performance of each subcarrier and the overall performance of a single link. The effects of interferences plays a detrimental role on the overall BER, and therefore increasing the SINR at each subcarrier can improve the overall BER. From a system level point of view, COFDM applied in a single user only mitigates the SNR fluctuations over different subcarriers of the same user, but cannot optimize the allocation of resources in a multiuser environment, such that the effect of interferences is minimized. Consequently, applying beamforming to each subcarrier can improve the performance of the system. Moreover, by beamforming at each subcarrier, we would be able to decrease the power consumption for achieving the same performance. The SINR fluctuation is amplified by the spatial processing and therefore the dependence of the overall BER on the SINR of each subcarrier becomes more severe. Power control can compensate for this fluctuation.

In this paper, we will consider four COFDM systems using TCM. Those are a COFDM system with no power control or beamforming, a COFDM system using the frequency-domain beamforming to increase the SINR with the same amount of total network power, a system with joint power control and frequency-domain beamforming per subcarrier, and finally a COFDM system where power control is performed per user and beamforming per subcarrier. In the second and third systems, the per subcarrier SINR is measured at the symbol level, before decoding (or demodulation). In the last system, the following iterative algorithm is used to achieve power control per user and beamforming per subcarrier. This algorithm tries to adapt the total user power by the equivalent SINR of the COFDM system derived from the BER of the receiver. The equivalent SINR of the COFDM system is defined as the SINR of an uncoded-OFDM system achieving the same BER, minus the coding gain of the code.

1) At step $n = 0$, all mobiles start with equal powers at all subcarriers. The weight vectors are initialized to a vector that have only one arbitrary nonzero components.
2) Each base station calculates the BER using a fixed number of frames.
3) Each base station calculates the SINR of the equivalent uncoded system using the relationship
$$\mathrm{BER} \simeq \frac{2}{k} Q\left(\sqrt{2\gamma_{\mathrm{uncoded}}}\sin\left(\frac{\pi}{M}\right)\right) \quad (20)$$
where $M = 2^k$ is the constellation size, and $\gamma_{\mathrm{uncoded}}$ is the SINR of the equivalent uncoded system [19]. This statement is an approximation of the bit error probability of AWGN channels for MPSK modulations. We have used it because the channel is assumed to be known at the receiver, and also the interference can be considered Gaussian, using central limit theorem.
4) The equivalent SINR of the COFDM system is calculated as $\gamma_{\mathrm{coded}} = \gamma_{\mathrm{uncoded}} - C$, where C is the coding gain of the coding scheme and the SINRs are evaluated in decibels.
5) The following relationship is used to calculate the total power of each mobile that is distributed equally among subcarriers:
$$P(n+1) = P(n)\frac{\gamma_{\mathrm{desired}}}{\gamma_{\mathrm{coded}}}. \quad (21)$$
6) Each base station uses (3) to find the beamforming weight at each subcarrier.
7) The algorithm is repeated until convergence.

Spatial processing improves the SINR in each subcarrier and the amount of improvement depends on the channel response, spatial signature and the interference in each subcarrier. Since some subcarriers get more benefits from the spatial processing, the power control per subcarrier saves more power compared with a system where the power control is performed per user.

VII. Simulation Results

We use a wireless network consisting of 36 base stations placed in a hexagonal pattern, where each cell contains one mobile. This model can be used by adopting any multiple-access scheme to distinguish the mobiles in a cell. We assume that all of the base stations belong to the same co-channel set. The formulations presented in this paper are applicable to the cases where multiple users are assigned to one base station, simply by allocating different indices to the same base stations associated with different mobiles. Users are randomly distributed in a cell according to a uniform distribution. We use an OFDM system with 32 subcarriers for transmission. The communication channel is assumed to follow the COST207 Typical Urban 6-ray channel model with average path delays of {0.0,0.2,0.5,1.6,2.3,5.0} measured in μs and path fading powers of {0.189, 0.379,0.239,0.095,0.061,0.037} [20]. The maximum channel delay spread is 5 μs and so the channel coherence bandwidth is 200 kHz. Path loss exponent is assumed to be four. Link loss also includes a shadow fading with 2.5 dB variance. We assume a quasi-static channel where the channel is assumed to be fixed over multiple OFDM symbols. Note that the subcarrier link gains for each user could be correlated and are obtained according to (2). The Doppler shifts are ignored, and any frequency and phase error and time mismatch are assumed to be resolved. Each OFDM symbol is assumed to be 1 μs long, which corresponds to a bandwidth of 1 MHz. Therefore, the subcarrier spacing is 32 kHz, which is smaller than the coherence bandwidth of the channel, and so the fading at each subcarrier can be considered flat. A one-tap frequency-domain equalizer is assumed at the receiver to compensate the channel flat fading at each subcarrier. The average power of the signal at each subcarrier at each transmitter is assumed to be unity. Noise power at each receiver is assumed to be -60 dBm. Quaternary phase-shift keying (QPSK) modulator and demodulator are used at all of the subcarriers in the transmitter and receiver. The desired SINR at each subcarrier for uncoded systems varies over a range of -5 to 15 dB. For the systems using beamforming a four-element antenna array is deployed at each receiver. In adaptive power policy, we measure the performance by picking a random mobile, and the selection is not important since the SINR at all of the subcarriers for all of the mobiles are almost the same. However, in uniform transmit power policy, the SINR at different receivers and subcarriers are different. Therefore we pick several base stations based on the average SINRs of their subcarriers to calculate the performances (e.g. the base stations labeled as "best base," "worst base," "base 0," and "average base" in Figs. 2 and 3).

In Fig. 2, a single-antenna configuration is used to perform adaptive and uniform power policy, when the channel is assumed to change from one OFDM symbol to another, but fixed during one OFDM symbol, and the total network power in the uniform policy experiment is the same as in the adaptive one. It is clear that the BERs of all of the base stations for the adaptive case are close to that of the base station having average SINR. The BER versus total network power is plotted in Fig. 3, when the channel follows a quasi-static model. In other words, in Fig. 3 the channel is assumed to be constant between two successive power updates, but it could vary from one to another, while in Fig. 2 the channel varies symbol by symbol.

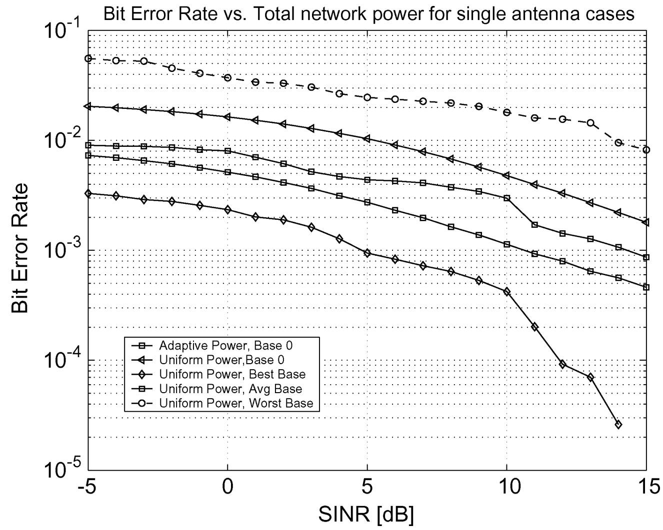

Fig. 2. BER versus SINR (in decibels) for single-antenna cases for time-varying channel between power updates.

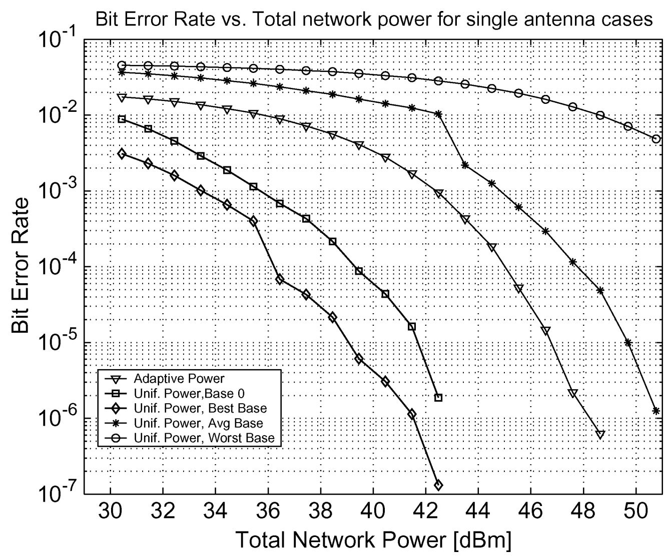

Fig. 3. BER versus total network power (in decibels relative to 1mW) for single-antenna cases assuming quasi-static channel.

We expect that in adaptive power policy all of the subcarriers perform close to the target SINR, while in uniform policy, because of different link gains at different subcarriers, the SINRs are expected to be different. This fact is illustrated in Fig. 4, where the SINRs for different subcarriers of a base station using both policies are shown. The target SNIR in this experiment is 5 dB. The sixth subcarrier is in deep fade, and therefore the transmitter power for this subcarrier is saturated at its maximum value, $P_{\max}$, and the target SINR is not achievable. The average SINR for the mobile chosen in uniform policy is about 5.1 dB.

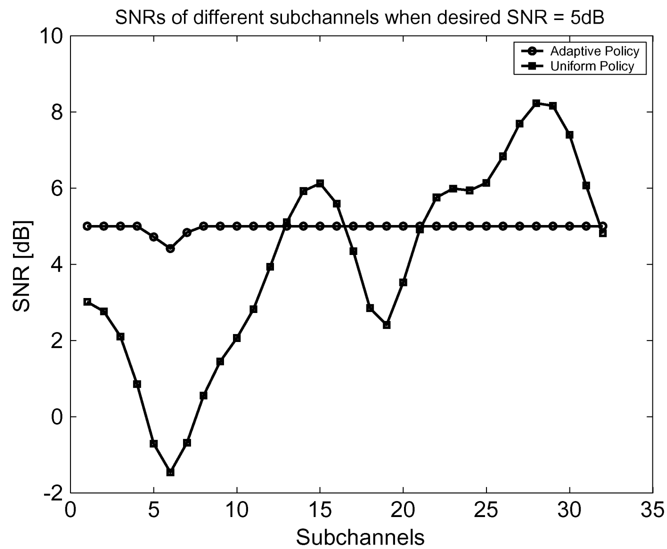

Fig. 4. SINRs of different subcarriers in the single-antenna cases.

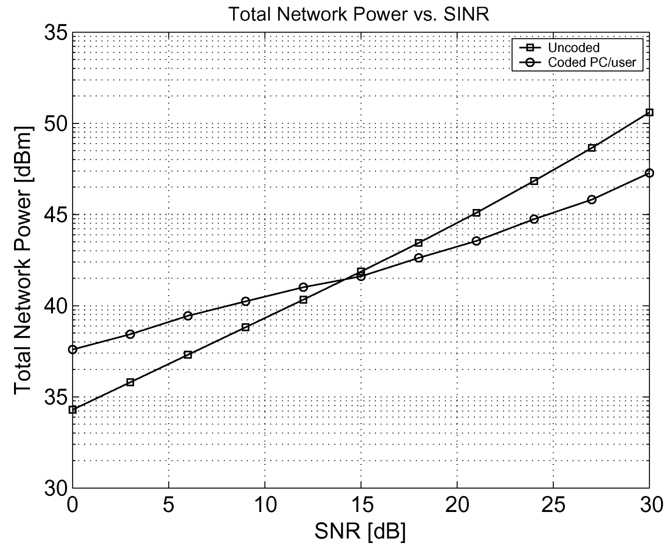

Fig. 6. Total network power verus desired SINR for coded and uncoded OFDM.

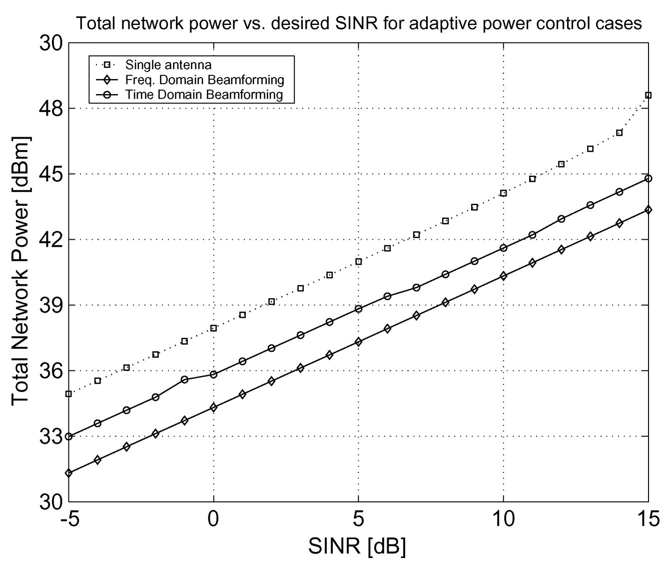

Fig. 5. Total network power (dBm) versus desired SINR (dB) for adaptive power control cases.

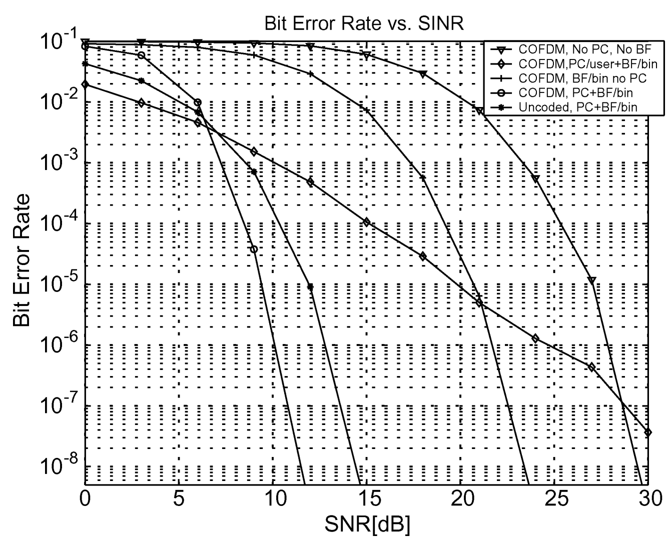

Fig. 7. BER versus desired SINR for coded and uncoded OFDM.

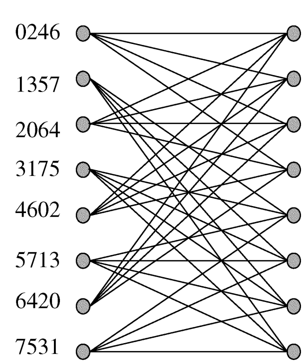

Fig. 8. 8-state 8-PSK TCM trellis.

Fig. 5 compares the three adaptive power control methods proposed in this paper. This figure shows that by using frequency-domain beamforming, we can achieve lower total network power for the same target SINR. For example, the 10 dB threshold SINR is achieved by reducing about 4 dB in total network power [the absolute powers are represented in dBm, but the differences are measured in decibels (dB)] compared with the single-antenna case. It is also shown that with the channel parameters we have used, the time-domain beamforming, although not optimal, performs somewhere between the single-antenna system and the system utilizing the frequency-domain beamforming at each receiver. In this case, for the same target SINR the total network power is about 3 dB lower compared with the single-antenna case. This amount clearly depends on the channel parameters. Obviously, the saving in complexity would be at the expense of performance. In all of these cases, the uncoded-OFDM is used and by using the adaptive power control scheme, we have guaranteed the SINR at all of the subcarriers to be close to the desired SINR. Since we have assumed a fixed modulation scheme at all of the subcarriers, we expect to achieve similar BERs in all of these cases. The simulation results have confirmed our expectation.

Figs. 6 and 7 compare the uncoded-OFDM system with the rate 2/3 COFDM system using the TCM represented in Fig. 8.

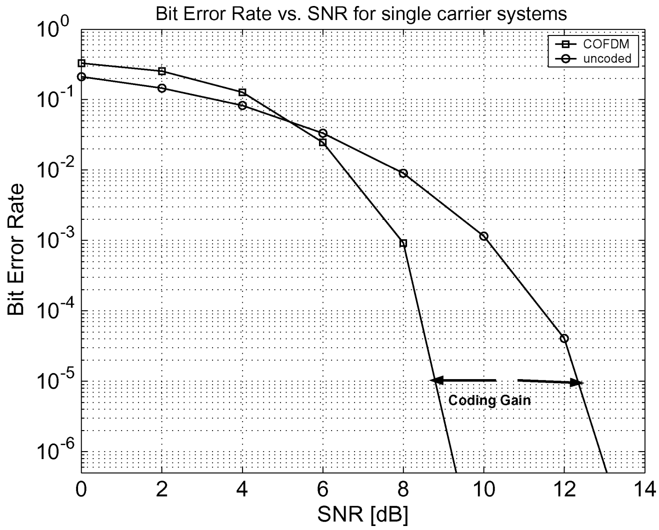

Fig. 9. Coding gain of the TCM encoder depicted in Fig. 8.

This is an Ungerboeck 8-state 8-PSK TCM encoder, whose minimum free distance d_{free}^2 is equal to 4.568 (no parallel transition) and the asymptotic coding gain (the coding gain at high SNR) is $\gamma = 2.29$ (3.6 dB). Viterbi decoding is used at each receiver. The equivalent uncoded system uses QPSK modulation. The SINR range for comparing the COFDM systems is chosen to be 0 to 30 dB. Fig. 9 is used to evaluate the coding gain at different SINRs. This figure is obtained by calculating the performance of a single carrier system using the same TCM encoder and Viterbi decoder (the arrows show the coding gain at BER $= 10^{-5}$). In Fig. 6, the total network power of an uncoded-OFDM system is compared with a COFDM system with per user power control and per subcarrier beamforming. Note that the total network power of a COFDM system with per subcarrier power control and beamforming is the same as that of the uncoded-OFDM system with power control and beamforming per subcarrier. The total power for uncoded system is lower than COFDM system for low or moderate SNR, but is higher for high SNR. This is compensated by lower BER shown in Fig. 7, where we compare the BER versus desired SINR for different OFDM systems. As can be seen from these curves, the COFDM system without any power control and beamforming has the lowest performance compared with other systems. A COFDM system where the transmitted powers are equal at all subcarriers but the frequency-domain beamforming is performed at each subcarrier, has a better performance compared with a COFDM system with no power control or beamforming. The curve marked by diamonds shows the BER of a COFDM system in which the per user power control jointly with per subcarrier frequency-domain beamforming (the algorithm mentioned in Section VI) is performed. This figure shows that if the joint power control and frequency-domain beamforming is performed at each subcarrier, both the uncoded and coded system have better performances compared with other configurations. For low SINR environments, the uncoded system achieves lower BER, while the performance of the coded system is better for the moderate and high SINR environments. As can be seen, these two curves intersects when the desired per subcarrier SINR is 7 dB. As the SINR is increased the coded system performs better. For low BERs, the OFDM coding gain is about 3.6 dB, which is consistent with the asymptotic coding gain of the trellis depicted in Fig. 8. Since in both cases power control and frequency-domain beamforming is performed at the symbol level, we expect the uncoded system to have a better performance in low SINRs, while in moderate or high SNR the coded system performs better.

VIII. Conclusion

We considered iterative joint power control and beamforming for OFDM networks with fixed modulation per subcarriers. Our study showed that the SINR at all subcarriers of all mobiles could be at least equal to a target value, while the total network power is minimized.

From our simulations we observed that by using joint frequency-domain beamforming and adaptive power control, we could achieve about 4 dB less total network power compared with the single-antenna case with the same target SINR. To reduce the complexity of the OFDM receivers, we performed the array processing in the time domain and provided an iterative algorithm to distribute the power among subcarriers. We observed that the performance in this case could be close to frequency-domain beamforming, while the complexity of the receiver is about 64 times less (for 128 subcarriers and 4 antennas). This reduction in complexity is achieved by paying a price of having higher total network power for the same target BER.

For some practical situation where array responses are unknown, MMSE time-domain beamforming jointly with power control was proposed. We have shown that the MMSE time-domain beamforming solution has the same SINR as the MVDR solution.

The proposed joint power control and frequency-domain beamforming was applied to the COFDM systems. We observed that an uncoded-OFDM system with the proposed algorithms performs better than the simple COFDM system, a coded system with per subcarrier beamforming with equal powers across subcarriers and a COFDM system with per user power control and per subcarrier beamforming. If the proposed algorithm is applied to COFDM, the BER is improved for moderate and high SINRs.

References

[1] J. T. E. McDonnel and T. A. Wilkinson, "Comparison of computational complexity of adaptive equalization and OFDM for indoor wireless networks," in *Proc. Pers., Indoor, Mobile Radio Commun. Conf.*, 1996, pp. 1088–1091.

[2] Y. Wu and W. Y. Zou, "Orthogonal frequency division multiplexing: A multicarrier modulation scheme," *IEEE Trans. Consum. Electron.*, vol. 41, pp. 392–398, Jun. 1995.

[3] J. A. C. Bingham, "Multicarrier modulation for data transmission: An idea whose time has come," *IEEE Commun. Mag.*, pp. 5–14, May 1990.

[4] A. Czylwik, "Adaptive OFDM for wideband radio channels," in *Proc. IEEE GlobeCom*, vol. 1, 1996, pp. 713–718.

[5] *Radio broadcast systems: Digital audio broadcasting (DAB) to mobile, portable and fixed receivers*, ETS 300-401, 1994.

[6] S. K. Lai, R. S. Cheng, K. Letaief, and R. D. Murch, "Adaptive trellis coded MQAM and power optimization for OFDM transmission," in *Proc. IEEE Veh. Technol. Conf.*, vol. 1, May 1999, pp. 290–294.

[7] H. Sari, G. Karam, and I. Jeanclaude, "Transmission techniques for digital terrestrial TV broadcasting," *IEEE Commun. Mag.*, pp. 100–109, Feb. 1995.

[8] J. A. C. Bingham, J. M. Cioffi, and P. S. Chow, "A practical discrete multitone tranceiver loading algorithm for data transmission over spectrally shaped channels," *IEEE Trans. Commun.*, vol. 43, pp. 773–775, Feb. 1995.

[9] R. F. H. Fischer and J. B. Huber, "A new loading algorithm for discrete multitone transmission," in *Proc. IEEE GlobeCom*, Feb. 1996, pp. 724–728.

[10] D. Hughes-Hartogs, "Ensembled modem structure for imperfect transmission media," U.S. Patent 4 679 227, Jul. 7, 1987.

[11] F. R. Farrokhi, L. Tassiulas, and K. J. R. Liu, "Joint optimal power control and beamforming in wireless networks using antenna arrays," *IEEE Trans. Commun.*, vol. 46, pp. 1313–1324, Oct. 1998.

[12] F. R. Farrokhi, K. J. R. Liu, and L. Tassiulas, "Transmit beamforming and power control for cellular wireless systems," *IEEE J. Sel. Areas Commun.*, vol. 16, pp. 1437–1450, Oct. 1998.

[13] G. J. Foschini, "A simple distributed autonomous power control algorithm and its convergence," in *Proc. IEEE Veh. Technol. Conf.*, vol. 42, Nov. 1993, pp. 641–646.

[14] S. Haykin, *Adaptive Filter Theory*, 3rd ed. Upper Saddle River, NJ: Prentice-Hall, 1996.

[15] F. R. Gantmacher, *The Theory of Matrices*, 3rd ed. New York: Chelsea, 1990.

[16] R. A. Monzingo and T. Miller, *Introduction to Adaptive Arrays*. New York: Wiley, 1980.

[17] G. Ungerboeck, "Trellis coded modulation with redundant signal set," *IEEE Commun. Mag.*, vol. 27, pp. 5–21, Feb. 1987.

[18] ——, "The state of the art in trellis coded modulation," in *Proc. 5th Tirrenia Int. Workshop Digital Commun.*, Tirrenia, Italy, Sep. 1991.

[19] J. Proakis, *Digital Communications*. Upper Saddle River, NJ: Prentice-Hall, 1995.

[20] A. Mehrotra, *GSM System Engineering*. Boston, MA: Artech House, 1996.

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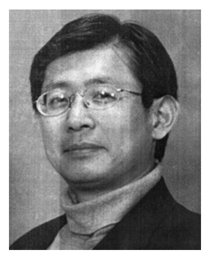

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