DOWNLINK BEAMFORMING FOR CELLULAR MOBILE COMMUNICATIONS

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ABSTRACT

A new technique for downlink transmission beamformer design in cellular mobile communications systems using an antenna array at the basestation is presented. The method is based on estimation of an underlying spatial distribution associated with each source's spatial downlink channel. The algorithm is "blind" in the sense that it depends only on uplink spatial channel statistics, requiring no mobile-to-basestation feedback in the design procedure. The assumed underlying spatial distribution models are general enough to be used in a wide variety of mobile communications scenarios (e.g., rural, urban, sub-urban, indoor). Simulation results verify the effectiveness of the new approach.

1. INTRODUCTION

Expected demand for mobile communications services is such that the use of *spatial diversity* to further improve spectral efficiency has recently received considerable attention [1]-[3]. Specifically, the use of antenna arrays in combination with signal processing algorithms at the basestation offer the possibility of exploiting spatial diversity present in the scenario to increase system capacity. In principle, these capacity enhancement strategies can be implemented in both uplink (mobile-to-basestation) and downlink (basestation-to-mobile) communication. Specifically, a set of weights is applied to the antenna array so as to reduce received (transmitted) co-channel interference in the uplink (downlink). The choice of weights is a function of the "spatial channel" formed between each co-channel mobile and each antenna element.

In the uplink, a training sequence can be used to design the weights according to a least mean squared error (LMSE) criterion as in [1]. This approach, known as "optimum combining," is applicable in a wide variety of mobile communication scenarios: indoor, urban, and sub-urban and rural. Generally, downlink weight design is more complicated. This is especially true in Frequency Division Duplex (FDD) systems where uplink and downlink communication take place at different frequencies. Thus, only if changes in the spatial channel are small over the time from start of the uplink frame to the end of the downlink frame *and* over the uplink-downlink frequency difference, will the downlink and uplink channel be approximately the same. Only then can the uplink antenna weights be used in the downlink. In practice (especially in FDD), uplink and downlink channel differences are often so large that the uplink weights cannot be used directly in the downlink.

In this paper, we address this problem by presenting a new technique for downlink transmission beamformer design which is especially appropriate for FDD systems such as GSM-900, DCS-1800 and PCS-1900. Unlike [3], the method is useful for cases where mobile-to-basestation feedback in the downlink beamformer design procedure is undesirable or not possible. The approach is similar in spirit to that presented in [2] wherein maximum likelihood estimates of a Gaussian parameterization of the spatial density of the users are employed. The method in this paper uses a least squares estimator for parameters of a more general Fourier based densities similar to that used in [4] in the context of array sensor noise modeling. This results in a computationally efficient algorithm which is appropriate for a wide variety of environmental scenarios.

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2. CELLULAR NETWORK STRUCTURE

Consider a network of clusters each containing C adjacent hexagonal cells. The cell radius is denoted as R. Each cell is further divided into Q sectors of width $\Delta = 2\pi/Q$. Antenna arrays (one per cell sector) in conjunction with appropriate signal processing techniques can increase system capacity by (i) reducing the channel re-use distance D by using fewer cells per cluster and/or (ii) by permitting multiple co-channel users *within* a sector. In this context, we focus on the problem of designing the downlink transmission beamformer to enhance downlink system capacity in the difficult yet common situation where the downlink channel cannot be estimated.

3. DATA AND CHANNEL MODELS

Consider a cell sector with an array of M elements. For some given uplink slot and some given uplink car-

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rier frequency, f_u, let N_x denote the number of received desired signals-of-interest (SOI's) due to the co-channel mobiles with uplink carrier frequency, f_u, located in the sector serviced by the array in question. Also, let N_i denote the number of received interfering signals-not-of-interest (SNOI's) due to co-channel mobiles at the same uplink carrier frequency, located in other sectors. Let $N = N_x + N_i$ denote the total number of received signals. The array snapshot $\mathbf{y}_u(f_u, t) = [y_{u_1}(f_u, t), \cdots, y_{u_M}(f_u, t)]^T$, can be modeled as:

$$\mathbf{y}_u(f_u, t) = \mathbf{A}_u(f_u, t)\mathbf{s}_u(f_u, t) + \mathbf{n}_u(f_u, t) \qquad (1)$$

The $M \times N$ matrix $\mathbf{A}_u(f_u, t)$ contains the time varying M dimensional spatial signature vectors, $\{\mathbf{a}_{u_k}(f_u, t)\}_{k=1}^{N}$ describing the uplink channels formed between each of the co-channel mobiles and the M antenna elements at carrier frequency f_u and time t. (Note that flat fading has been assumed.) The N dimensional vector $\mathbf{s}_u(f_u, t)$ contains the signals $\{s_{u_k}(f_u, t)\}_{k=1}^{N}$ transmitted by the N co-channel mobiles at carrier frequency f_u during the slot in question. The vector of additive sensor noise is denoted as $\mathbf{n}_u(f_u, t)$.

Analogously, consider the same uplink co-channel users now receiving the basestation transmission during the corresponding downlink slot at carrier frequency f_d. We can group the received SOI and SNOI signals into an N dimensional user downlink "snapshot" vector, $\hat{\mathbf{s}}_d(f_d, t)$:

$$\hat{\mathbf{s}}_d(f_d, t) = \mathbf{A}_d^T(f_d, t)\mathbf{y}_d(f_d, t) + \mathbf{n}_d(f_d, t) \qquad (2)$$

where $(\cdot)^T$ denotes matrix/vector transpose, and $\mathbf{A}_d(f_d, t)$, $\mathbf{y}_d(f_d, t)$, and $\mathbf{n}_d(f_d, t)$ are respectively, the spatial signature matrix, the antenna element transmission vector, and the additive Gaussian noise vector at the mobiles.

Let us consider one of the co-channel users, user k, in the given uplink and downlink slots and model its corresponding uplink and downlink channels for all time t. To model frequency hopping, the associated carrier frequency is indexed by the particular mobile under consideration. This user's uplink spatial signature at time t can be written as a weighted average of point source signatures:

$$\mathbf{a}_{u_k}(f_{u_k}, t) = \int_{\theta \in \Theta} \mathbf{v}(\theta | f_{u_k}) g_{u_k}(\theta | f_{u_k}, t) d\theta \qquad (3)$$

$$\mathbf{v}(\theta | f_{u_k}) = \left[1, e^{j2\pi f_{u_k}\frac{z}{c}\sin\theta}, \cdots, e^{j(M-1)2\pi f_{u_k}\frac{z}{c}\sin\theta}\right]^T \qquad (4)$$

where $\mathbf{v}(\theta | f_{u_k})$ is the standard far-field, narrow band point source steering vector associated with the uniform linear array (ULA), θ is the angle of incidence (with respect to the array broadside), z is the interelement antenna spacing, c is the propagation speed and $g_{u_k}(\theta | f_{u_k}, t)$ is a "spatial weighting function." The interval over which integration is carried out, Θ is the array's angular coverage interval and will depend on the directionality of the antenna elements (which, in turn, is largely determined by the cell *sectorization* scheme employed.)

In general, the uplink weighting function for the kth user, $g_{u_k}(\theta | f_{u_k}, t)$ can be written as:

$$g_{u_k}(\theta | f_{u_k}, t) = \sqrt{\frac{L_k(t)}{D_k^{\gamma}(t)}} \beta_{u_k}(\theta | f_{u_k}, t) e^{j\alpha_{u_k}(\theta | f_{u_k}, t)} p_{u_k}(\theta | t) \qquad (5)$$

where $L_k(t)$ is a zero mean log-normally distributed shadowing term with $E([10 \log L_k(t)]) = 0$ and $E([10 \log L_k(t)]^2) = \sigma_L^2$ for the kth user. The path loss term is denoted as $\sqrt{1/D_k^{\gamma}(t)}$ where $D_k(t)$ and γ are the distance between the kth user and the basestation (normalized by the cell radius, R), and the path loss exponent, respectively. The unit variance Rayleigh distributed gain function and the uniformly distributed phase function are denoted as $\beta_{u_k}(\theta | f_{u_k}, t)$ and $\alpha_{u_k}(\theta | f_{u_k}, t)$, respectively. Their product models the fast fading component of the channel. Lastly, $p_{u_k}^2(\theta | t)$ is the non-negative, unit area underlying "spatial density function" associated with user k. In practice, the ray gain and phase functions can be expected to change far more rapidly with source movement than the spatial density function.

The weighting function $g_{u_k}(\theta | f_{u_k}, t)$ will be a zero mean random function of angle conditioned on frequency and time and of correlation: $r_{u_k}(\theta | f_{u_k}, f_{u_k}(t'), t, t')\delta(\theta - \theta')$ where $\delta(\cdot)$ denotes the Dirac delta function, and it has been recognized that the channel at one angle of arrival is uncorrelated with that at other angles of arrival.

Analogously, the downlink spatial signature vector for user k at time t can be expressed as:

$$\mathbf{a}_{d_k}(f_{d_k}, t) = \int_{\theta \in \Theta} \mathbf{v}(\theta | f_{d_k}) g_{d_k}(\theta | f_{d_k}, t) d\theta \qquad (6)$$

$$g_{d_k}(\theta | f_{d_k}, t) = \sqrt{\frac{L_k(t)}{D_k^{\gamma}(t)}} \beta_{d_k}(\theta | f_{d_k}, t) e^{j\alpha_{d_k}(\theta | f_{d_k}, t)} p_{d_k}(\theta | t)$$

4. DOWNLINK OPTIMUM COMBINING

In the downlink, the goal is that of designing a set of N_x M-dimensional weight vectors $\{\mathbf{w}_{d_k}(f_{d_k}, t)\}_{k=1}^{N_x}$ which when weighted by the (assumed unit power) transmitted SOI signals, $\{s_{d_k}(f_{d_k}, t)\}_{k=1}^{N_x}$ give rise to:

$$\mathbf{a}_{d_k}^H(f_{d_k}, t)\mathbf{y}_d(f_{d_k}, t) \approx s_{d_k}(f_{d_k}, t), \qquad k \in \{1, \cdots, N_x\}$$

$$\mathbf{a}_{d_k}^H(f_{d_k}, t)\mathbf{y}_d(f_{d_k}, t) \approx 0, \qquad k \in \{N_x + 1, \cdots, N\}$$

$$\mathbf{y}_d(f_{d_k}, t) = \sum_{k=1}^{N_x} \mathbf{w}_{d_k}^*(f_{d_k}, t) s_{d_k}(f_{d_k}, t) \qquad (7)$$

where $(\cdot)^*$ denotes complex conjugation.

Each weight vector can be designed such that the total associated interference power is minimized subject to a desired mobile received power constraint:

$$\mathbf{w}_{d_k} = \arg\min_{\mathbf{w}} \mathbf{w}^H \mathbf{R}_{d_k}^{[i]} \mathbf{w}, \;\; \mathbf{w}^H \mathbf{R}_{d_k}^{[s]} \mathbf{w} = \chi_k = \frac{\mathrm{tr}\left(\mathbf{R}_{u_k}^{[s]}\right)}{M \zeta_{u_k}} \qquad (8)$$

$$\mathbf{R}_{d_k}^{[s]} = \mathbf{R}_{d_k}, \qquad \mathbf{R}_{d_k}^{[i]} = \sum_{q=1, q\neq k}^{N} \mathbf{R}_{d_q}, \quad \mathbf{R}_{d_q} = \mathbf{a}_{d_q}\mathbf{a}_{d_q}^H,$$

where $\mathbf{R}_{d_k}^{[s]}$ and $\mathbf{R}_{d_k}^{[i]}$ are respectively the downlink signal and interference correlation matrices associated with the kth user, and $\mathrm{tr}(\cdot)$ denotes the matrix trace operation. Note the constraint in (8) requires that the received power at the desired user equal the mean power received by each antenna from this user in the uplink normalized by the mobile's transmission power ζ_{u_k}. This type of constraint is preferred to an absolute power type constraint such as $\mathbf{w}^H\mathbf{R}_{d_k}^{[s]}\mathbf{w} = 1$ because the latter may place very high attenuation requirements on the design of the downlink beamformer. For example, consider two co-channel users the norms of whose uplink channel vectors differ, say, by 30dB, due to path loss differences, etc. The constraint $\mathbf{w}^H\mathbf{R}_{d_k}^{[s]}\mathbf{w} = 1$ implies that transmission power directed toward the weak user (by the transmission beamformer for the weak user) will be 30dB larger than that directed toward the strong user (by the transmission beamformer for the strong user). This in turn implies that, in order to achieve a downlink signal-to-interference ratio (SIR) of say, 10dB at the strong user, the beamformer for the weak user must direct at least 40dB less transmission power toward the strong user relative to the weak user. On the other hand, if the constraint in (8) is used, then to achieve10dB SIR at the strong user, the beamformer for the weak user need only direct at least10dB less toward the strong user relative to the weak user.

The solution is proportional to $\mathbf{e}_{d_k}^{[\max]}$ the generalized eigenvector associated with the maximum generalized eigenvalue of the kth signal and interference and noise correlation matrix pair: $\{\mathbf{R}_{d_k}^{[s]}, \mathbf{R}_{d_k}^{[i]}\}$:

$$\mathbf{w}_{d_k} = \mathbf{e}_{d_k}^{[\max]}\sqrt{\chi_k \Big/ \mathbf{e}_{d_k}^{[\max]H}\mathbf{R}_{d_k}^{[s]}\mathbf{e}_{d_k}^{[\max]}} \tag{9}$$

5. ALTERNATIVE COMBINER DESIGN

The downlink combiner design proposed in (8) requires full knowledge of the downlink channels associated with the user of interest as well as all effected interfered mobiles. Often in practice, such information will *not* be available. In this section we propose an alternative downlink combiner for such cases.

The technique is most easily derived in the context of a Time Division Multiple Access (TDMA) system with the following assumptions: First, for each user the uplink and subsequent downlink channels are uncorrelated (in the sense that their random weighting functions in (3) and (6) are uncorrelated). Next, a *training sequence* of duration T_t is available in uplink slot for each user. Such sequences are incorporated into existing system standards. Also, the uplink channel for a user in a given frame is *uncorrelated* with the uplink channel for the same user in another frame. This will very much be the case in frequency hopping systems since the gain and phase fading functions of (5), $\beta_{u_k}(\theta|f_{u_k}, t)$ and $\alpha_{u_k}(\theta|f_{u_k}, t)$ are highly sensitive to changes in the carrier frequency. Lastly, the underlying spatial densities associated with the uplink and downlink channels, $p_{u_k}^2(\theta|t)$ and $p_{d_k}^2(\theta|t)$, of (5) and (7), respectively, are assumed to change *very slowly* with time compared to the gain and phase fading functions. In particular, we assume that these underlying spatial densities are approximately constant over several frames. This is reasonable since these functions do not exhibit great fluctuation in response to changes in the carrier frequency and/or "small" changes in the mobile position. Moreover, it is assumed that the underlying uplink and downlink spatial densities are approximately equal:

$$p_k^2(\theta|t) \approx p_{u_k}^2(\theta|t) \approx p_{d_k}^2(\theta|t). \tag{10}$$

This is the information which is assumed *common* to both the uplink and the downlink (in addition to the usually considered common log-normal fading and path loss components) and will form the basis of the combiner design procedure described below.

In such a case, a *modified* version of the optimum combiner based on parametric signal and interference and noise correlation matrices which are *averaged* over the fast fading terms can be formulated. Let us define $\overline{\mathbf{R}}_{d_k}^{[s]}$ as the downlink correlation matrix associated with the kth user which has been averaged over the fast fading:

$$\overline{\mathbf{R}}_{d_k}^{[s]} = E_{\beta,\alpha}[\mathbf{a}_{d_k}\mathbf{a}_{d_k}^H] = \int_{\theta\in\Theta} r_{d_k}(\theta)\mathbf{v}(\theta|f_{d_k})\mathbf{v}^H(\theta|f_{d_k})d\theta$$

where $r_{d_k}(\theta) = \frac{L_k}{D_k^\gamma}p_{d_k}^2(\theta)$ and $E_{\beta,\alpha}[\cdot]$ denotes expectation over the fast fading. Based on average correlation matrices of the form in (11), we can reformulate the constrained minimum interference power criterion of (8) (based on information that cannot be inferred from the uplink) as the following constrained minimum average interference power criterion (based on information that *can* be inferred from the uplink):

$$\overline{\mathbf{w}}_{d_k} = \arg\min_{\mathbf{w}} \mathbf{w}^H\overline{\mathbf{R}}_{d_k}^{[i]}\mathbf{w},\ \mathbf{w}^H\mathbf{R}_{d_k}^{[s]}\mathbf{w} = \bar{\chi}_k = \frac{\mathrm{tr}\left(\overline{\mathbf{R}}_{u_k}^{[s]}\right)}{M\zeta_{u_k}} \tag{11}$$

$$\overline{\mathbf{R}}_{d_k}^{[i]} = \sum_{q=1, q\neq k}^{N} \overline{\mathbf{R}}_{d_q},\ \overline{\mathbf{R}}_{d_k}^{[s]} = \overline{\mathbf{R}}_{d_k},\ \overline{\mathbf{R}}_{d_q} = E_{\beta,\alpha}\left[\mathbf{a}_{d_q}\mathbf{a}_{d_q}^H\right]$$

The solution is given as: $\overline{\mathbf{w}}_{d_k} = \overline{\mathbf{e}}_{d_k}^{[\max]}\sqrt{\bar{\chi}_k \Big/ \overline{\mathbf{e}}_{d_k}^{[\max]H}\overline{\mathbf{R}}_{d_k}^{[s]}\overline{\mathbf{e}}_{d_k}^{[\max]}}$, with $\overline{\mathbf{e}}_{d_k}^{[\max]}$ denoting the "maximum" generalized eigenvector of $\{\overline{\mathbf{R}}_{d_k}^{[s]}, \overline{\mathbf{R}}_{d_k}^{[i]}\}$.

The downlink combiner proposed in (11) is forced to enhance reception of the user of interest and attenuate the interferers on the basis of magnitude as a function of angle. The combiner will attempt to increase the magnitude of its spatial response in those directions where the desired user is underlying spatial density is large while trying to attenuate it in those directions where the interferers spatial density functions are large. Performance will depend on the extent to which the spatial density functions of the users overlap. While the above approach is clearly sub-optimal, it makes the best out of a difficult situation–fully exploiting all information about the downlink channel that can be obtained from the uplink.

6. IMPLEMENTATION

We now explain how estimates of the corresponding average downlink signal and interference correlation matrices, $\overline{\mathbf{R}}_{d_k}^{[s]}$ and $\overline{\mathbf{R}}_{d_k}^{[i]}$, respectively, are obtained for use in the new procedure. Since the uplink spatial weighting correlation function is defined only over the sector, it can be represented by a *Fourier series expansion* as first proposed in [4] for modeling noise statistics. The Fourier series expansion of the spatial weighting correlation function over the interval $\theta \in [-\Delta/2, \Delta/2)$ can be written as:

$$r_k(\theta) = \sum_{l=-\infty}^{\infty} c_{k_l} e^{jlQ\theta}, \; c_{k_l} = \frac{1}{\Delta}\int_{-\Delta/2}^{\Delta/2} r_k(\theta) e^{-jlQ\theta} d\theta \tag{12}$$

Now, since the correlation function will often be a quite smooth function of angle, in practice a truncated version of (12) will usually be a sufficient approximation.

$$r_k(\theta) \approx \textstyle\sum_{l=-L+1}^{L-1} c_{k_l} e^{jlQ\theta}, \quad \theta \in [-\Delta/2, \Delta/2) \tag{13}$$

The average uplink and downlink correlation matrices can be approximated as [4]:

$$\overline{\mathbf{R}}_{u_k}^{[s]} = E_{\beta,\alpha}\left[\mathbf{a}_{u_k}\mathbf{a}_{u_k}^H\right] \approx \sum_{l=-L+1}^{L-1} c_{k_l} \mathbf{\Sigma}_{u_k}^{[l]} \tag{14}$$

$$\overline{\mathbf{R}}_{d_k}^{[s]} = E_{\beta,\alpha}\left[\mathbf{a}_{d_k}\mathbf{a}_{d_k}^H\right] \approx \sum_{l=-L+1}^{L-1} c_{k_l} \mathbf{\Sigma}_{d_k}^{[l]} \tag{15}$$

$$\mathbf{c}_k = \left[c_{k_{-L+1}}, c_{k_{-L+2}}, \cdots, c_{k_{L-1}}\right]^T$$

$$\mathbf{\Sigma}_{u_k}^{[l]} = \int_{-\Delta/2}^{\Delta/2} \mathbf{v}(\theta|f_{u_k})\mathbf{v}^H(\theta|f_{u_k}) e^{jlQ\theta} d\theta$$

$$\mathbf{\Sigma}_{d_k}^{[l]} = \int_{-\Delta/2}^{\Delta/2} \mathbf{v}(\theta|f_{d_k})\mathbf{v}^H(\theta|f_{d_k}) e^{jlQ\theta} d\theta$$

Thus, the problem of estimating the spatial weighting correlation function from the uplink data can be posed as the problem of estimating the parameter vector $\mathbf{c}_k$ from the uplink data.

The spatial signature estimates are obtained by post-multiplying an uplink snapshot data matrix by the pseudoinverse of a known "training sequence signal matrix." In particular, if each uplink slot contains a training sequence of length T_t starting at time τ_t, the spatial signature estimate is calculated as:

$$\widehat{\mathbf{A}}_u(i) = [\hat{\mathbf{a}}_{u_1}(i), \hat{\mathbf{a}}_{u_2}(i), \cdots \hat{\mathbf{a}}_{N_x}(i)] = \mathbf{Y}_u(i)\mathbf{S}^{\#}(i) \tag{16}$$

$$\mathbf{Y}_u(i) = [\mathbf{y}_{u_k}(\tau_t + iT_{fu}) \cdots \mathbf{y}_{u_k}([J-1]T_o + \tau_t + iT_{fu})]$$

$$\mathbf{S}(i) = \begin{bmatrix} s_1(\tau_t + iT_{fu}) & \cdots & s_1([J-1]T_o + \tau_t + iT_{fu}) \\ \vdots & \cdots & \vdots \\ s_{N_x}(\tau_t + iT_{fu}) & \cdots & s_{N_x}([J-1]T_o + \tau_t + iT_{fu}) \end{bmatrix}$$

which is valid, without loss of generality for the first user slot in the uplink frame, and where T_o is the uplink sampling rate ($T_t = JT_o$) and $(\cdot)^{\#}$ denotes the matrix pseudo-inverse operation. (Note that it is assumed that the training sequences for each of the users are linearly independent.) The average uplink correlation matrices can be estimated by averaging the outer product of spatial signature estimates for a given user obtained over a number of previous frames.

$$\widehat{\overline{\mathbf{R}}}_{u_k} = \tfrac{1}{F}\textstyle\sum_{i=0}^{F-1} \hat{\mathbf{a}}_{u_k}(i)\hat{\mathbf{a}}_{u_k}^H(i). \tag{17}$$

Ideally, frequency hopping is performed over a band sufficiently wide so as to decorrelate fast fading from one frame to the next, but sufficiently narrow so as to produce negligible changes in the uplink steering vector (4) as a function of frequency. This further implies that the dependence of $\mathbf{\Sigma}_{u_k}^{[l]}$ on k can, in effect, be eliminated.

$$\mathbf{\Sigma}_{u_k}^{[l]} \approx \mathbf{\Sigma}_u^{[l]} = \textstyle\int_{-\Delta/2}^{\Delta/2} \mathbf{v}(\theta|\overline{f}_u)\mathbf{v}^H(\theta|\overline{f}_u) e^{jlQ\theta} d\theta \tag{18}$$

where $\overline{f}_u$ is a "nominal" uplink carrier frequency.

Now, returning to the estimation problem, one approach is that of a simple parametric least squares fit of the estimated uplink correlation matrix:

$$\hat{\mathbf{c}}_k = \arg\max_{\mathbf{c}_k} \left\| \widehat{\overline{\mathbf{R}}}_{u_k} - \textstyle\sum_{l=-L+1}^{L-1} c_{k_l} \mathbf{\Sigma}_u^{[l]} \right\|_F^2 \tag{19}$$

where $\|\cdot\|_F^2$ denotes the square of the matrix Frobenious norm. The solution is easily shown to be [5]:

$$\hat{\mathbf{c}}_k = \mathbf{G}^{-1}\mathbf{g}_k, \quad [\mathbf{G}]_{mn} = \operatorname{tr}\left(\mathbf{\Sigma}_u^{[m-L]}\mathbf{\Sigma}_u^{[n-L]}\right), \tag{20}$$

$$[\mathbf{g}_k]_m = \operatorname{tr}(\widehat{\overline{\mathbf{R}}}_{u_k}\mathbf{\Sigma}_u^{[m-L]}) \qquad m, n \in \{1, \cdots, 2L-1\}$$

where $\operatorname{tr}(\cdot)$, $[\cdot]_{mn}$, and $[\cdot]_m$ denote matrix trace, the mnth element of a matrix, and the mth element of a vector, respectively. Note that the above estimator is very efficient computationally since $\mathbf{G}^{-1}$ can be computed off-line. Once the Fourier parameter vectors are estimated, the associated parametric downlink correlation matrix estimates are simply formed as in (15):

$$\widehat{\overline{\mathbf{R}}}_{d_k}^{[i]} = \sum_{\substack{q=1 \\ q\neq k}}^{N} \widehat{\overline{\mathbf{R}}}_{d_q}, \quad \widehat{\overline{\mathbf{R}}}_{d_k}^{[s]} = \widehat{\overline{\mathbf{R}}}_{d_k}, \quad \widehat{\overline{\mathbf{R}}}_{d_q} \equiv \sum_{l=-L+1}^{L-1} c_{q_l} \mathbf{\Sigma}_{d_k}^{[l]},$$

and the weight vector is calculated as:

$$\widehat{\overline{\mathbf{w}}}_{d_k} = \hat{\overline{\mathbf{e}}}_{d_k}^{[\max]} \sqrt{\bar{\chi}_k \Big/ \hat{\overline{\mathbf{e}}}_{d_k}^{[\max]H} \widehat{\overline{\mathbf{R}}}_{d_k}^{[s]} \hat{\overline{\mathbf{e}}}_{d_k}^{[\max]}} \tag{21}$$

with $\hat{\overline{\mathbf{e}}}_{d_k}^{[\max]}$ denoting the "maximum" generalized eigenvector of $\{\widehat{\overline{\mathbf{R}}}_{d_k}^{[s]}, \widehat{\overline{\mathbf{R}}}_{d_k}^{[i]}\}$.

7. RESULTS

For the simulations, we consider the case where none of the neighboring cells which interfere with the sector in question uses antenna arrays. This simple case is interesting because it can be used to evaluate the *progressive* introduction of antenna arrays in existing cellular communications networks. It also represents a pessimistic, worst case out-of-sector interference scenario when the other cells do employ antenna arrays.

$C = 4$ cells per cluster and $Q = 3$ sectors per cell are considered. In addition to the desired transmitted signal, the mobile-of-interest (MOI) will receive interfering transmissions to $N_x - 1$ co-channel mobiles in the same sector as well as N_i transmissions to co-channel mobiles in near-by (first tier) clusters which are assumed to use single antenna transmission which is omnidirectional over the sector.

The underlying densities for the users are generated randomly. As in [2], $D_k = 0.5$, $\sigma_L = 6dB$, and $\gamma = 3.5$. While the proposed technique is appropriate for a wide variety of underlying spatial densities, the densities used in the simulation are synthesized as a sum of (in our case, two) Gaussian densities of uniformly distributed mean angle over the sector, Θ. Performance is measured in terms of average SINR at the MOI where the averaging is now carried out over the underlying spatial density functions of the co-sector users. 500 runs are considered for each average SINR calcuation.

Consider first the case of low angular dispersion–uniformly distributed standard deviation on the interval $[0, \pi/200)$. The resulting densities essentially consist of two point sources which may be an appropriate model for ray type multipath in rural scenarios. Fig. 1 shows average SINR as a function of the number of co-channel users in the sector for a single antenna system and the $M = 8$ element array using the proposed transmission beamforming technique. ($L = M$.) Substantial improvement over the single antenna case is seen for all number of users considered.

Next, attention is turned to the case of high angular dispersion in the angular densities–uniformly distributed standard deviation on the interval $[0, \pi/20)$. This model may be more appropriate for urban or indoor applications. Fig. 2 shows average SINR as a function of the number of co-channel users in the sector for a single antenna system and the eight element array using the proposed transmission beamforming technique. Again, substantial improvement over the single antenna case is seen for all number of users considered. However, there is a degradation in beamformer performance with respect to the to the low angular spreading case. This is to be expected since extra beamformer degrees of freedom are required to create broad nulls in the directions associated with the interfered users.

8. CONCLUSION

A new technique for downlink transmission beamformer design in cellular communications systems has been presented. The algorithm requires no mobile-to-base station feedback, is computationally efficient, and is well suited for a wide variety of scenarios typically found in mobile communications. As for future work, the method will be implemented in the ACTS 020 TSUNAMI (II) project field trials. Moreover, the technique can be easily extended to frequency-selective channels by using a two-dimensional series approximation of the corresponding spatio-temporal distribution.

9. REFERENCES

[1] J.H. Winters. Optimum combining in digital mobile radio with cochannel interference. *IEEE Journal on Selected Areas in Communications*, vol. 2: pp. 528–539, July 1984.

[2] P. Zetterberg and B. Ottersten. "The spectrum efficiency of a basestation antenna array system for spatially selective transmission". *IEEE Transactions on Vehicular Technology*, vol. 44: pp. 651–660, August 1995.

[3] D Gerlach and A. Paulraj. "Adaptive transmitting antenna arrays with feedback". *IEEE Signal Processing Letters*, vol. 1: pp. 150–152, October 1994.

[4] B. Friedlander and A. Weiss. "Direction finding using noise covariance modeling". *IEEE Transactions on Signal Processing*, vol. 43: pp. 1557–1567, July 1995.

[5] L.L. Scharf. *Statistical Signal Processing–Detection, Estimation, and Time Series Analysis.* Addison-Wesley, 1991.

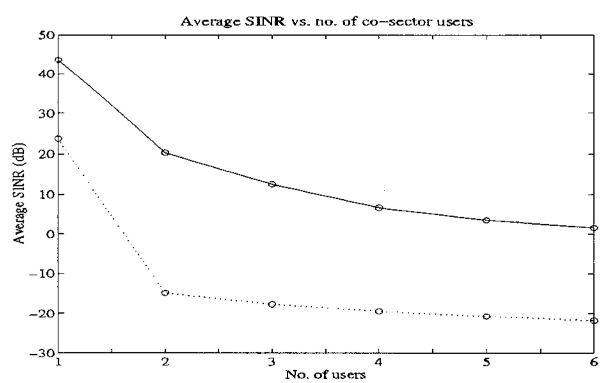

Figure 1: Low spreading; solid $M = 8$, dotted $M = 1$.

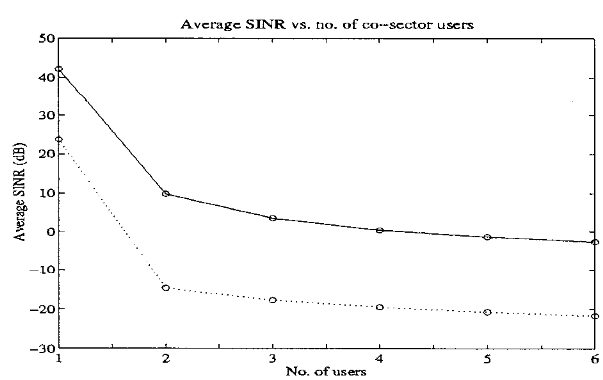

Figure 2: High spreading; solid $M = 8$, dotted $M = 1$.