A BLIND ADAPTIVE TRANSMIT ANTENNA ALGORITHM FOR WIRELESS COMMUNICATION

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ABSTRACT

A method is proposed for forming an adaptive phased array transmission beam pattern at a base station without any knowledge of array geometry or mobile feedback. Estimates of the receive vector propagation channels are to objectives and constraints in transmit weight vector optimization problems leading to a principal eigenvector solution in the single user case, or a principal generalized eigenvector solution in the multiple user case. We show through simulation of a multiple user cellular network that our algorithm is capable of improving network frequency re-use capacity by a factor of 3 to 6.

1. INTRODUCTION

THE problem we consider is that of forming transmission beam patterns at a wireless base station in a frequency division duplex (FDD) communication system where the transmit frequency is slightly offset from the receive frequency. There are three existing approaches to applying directive transmit antennas to improve FDD wireless communication link performance - fixed beam selection (sectoring), adaptive beam forming based on receive channel angle estimates [1], and adaptive transmission based on mobile feedback [2].

The approach proposed herein offers the advantages of adaptive transmission using feedback without the mobile radio complexity increase and information capacity penalty. Multipath propagation effects are explicitly accounted for in the theoretical approach to the problem. The technique is blind in that no explicit knowledge of the array geometry, array calibration, or mobile feedback is required. The receive and transmit antenna geometries must obey certain criteria but this criteria need not be known or accounted for by the transmit beam forming algorithm. The phase and amplitude differences the receiver and transmitter electronics must also be equalized. Recent results in blind signal copy of multiple co-channel signals using antenna arrays are exploited to make possible the estimation of the receive signal channel vector. The optimum transmit beam pattern is then found by solving a quadratic optimization problem based on receive channel covariance matrix estimates.

Single user beam forming is developed in Section 2. Section 3 extends the approach to a multiple user scenario. Section 4 presents multiple co-channel user simulation results. Conclusions are offered in section 5.

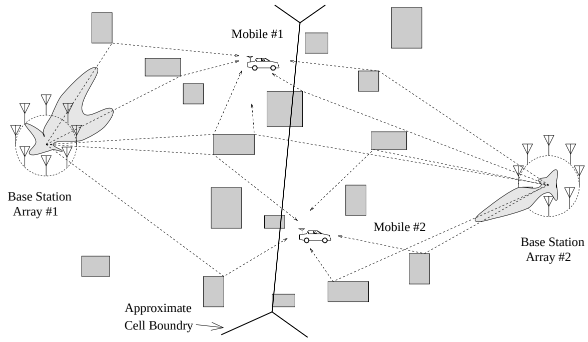


Figure 1: An illustration of the propagation channel

2. SINGLE USER BEAM FORMING

In an FDD wireless communication link, information about the base station transmit channel is usually unavailable to the base station. This motivates us to use receive channel information to form the transmit beam. We propose that while there is no correlation between the instantaneous values of the transmit and receive vector channels, there is strong correlation between the *average* receive channel vector subspace and the *average* transmit channel vector subspace. There are two fundamental observations which justify this proposition:

- *The propagation physics of the wireless vector channel are such that certain short-term time average statistics are invariant to small frequency translations.*
- *If the transmit and receive antenna geometries are properly chosen, the array response due to a given path angle is also largely invariant to small frequency translations.*

2.1. Transmit Channel Model

We assume that the channel delay spread and the signal bandwidth are related by $B_s \ll \frac{1}{\tau_{max}}$. This simplifies the discussion. Extensions to the delay spread channel will be reported in a future paper.

An illustration of the propagation channel is presented in Figure 2.1. Only azimuth angles are considered in the propagation geometry but the results can be generalized to three dimensions. The mobile is assumed to have a single omni-directional antenna while the base station has M omni-directional elements. Energy radiated from each base station antenna element is reflected by large dominant reflectors. Many local scatterers in the near vicinity of the mobile again reflect the energy before it finally reaches the mobile. As the vehicle position varies, the phase length to each local scatterer changes resulting in a time variant vector propagation channel from the base station transmit array to the mobile [3].

The vector channel model for the signal received by the mobile user is

$$\begin{aligned} y_R(t) &= \mathbf{w}^H(t)\left[\sum_{l=1}^{L}\alpha_{T,l}(t)\mathbf{a}_T(\theta_l)s_T(t)\right]+n_T(t) \\ &= \mathbf{w}^H(t)\mathbf{a}_T(t)s_T(t)+n_T(t) \end{aligned} \quad (1)$$

where the complex transmitted signal is given by $s_T(t)$. The M-element complex row vector $\mathbf{w}^H(t)$ is the transmitted array beam pattern. The time varying path amplitude $\alpha_{T,l}(t)$ is the complex path loss for the l^{th} dominant path. The array response $\mathbf{a}_T(\theta_l)$ to the angle θ_l is determined by the geometry of the array and is characterized by an array manifold mapping of path angle to M-element complex vector [4]. The total number of paths which contribute significantly to the channel is L.

2.2. Optimum Single User Beam Forming

Our objective is to determine a beam pattern which maximizes the signal to noise ratio for $y_R(t)$ subject to a total radiated power constraint $|\mathbf{w}(t)|^2 \le P_{T_{max}}$. For an FDD radio link, where transmit to receive frequency separation is greater than the channel coherence bandwidth [5], the instantaneous transmit channel is uncorrelated from the receive channel. Therefore, we choose to address the problem of delivering the largest possible *average* power to the mobile receiver subject to certain constraints. The time dependence will be dropped from the weight vector notation. This does not imply that the weight vector will be constant for all time. This does imply that the weight vector will adapt to slowly varying components in the channel variation and will not track channel components which vary significantly over the time periods used to collect the required channel statistics.

We define the optimum average power transmit weight vector as the beam pattern vector which maximizes the average signal power received by the desired mobile, subject to a maximum radiated power constraint.

$$\begin{aligned} \mathbf{w}_{opt} &= arg\left(\max_{\mathbf{w}}\left[E\left(|\mathbf{w}^H\mathbf{a}_T(t)s_T(t)|^2\right)\right]\right) \\ &= arg\left(\max_{\mathbf{w}}\left[\mathbf{w}^H\mathbf{R}_T\mathbf{w}\right]\right) \\ & s.t. \\ \|\mathbf{w}\|_2^2 &= P_{T_{max}} \end{aligned} \quad (2)$$

where $\mathbf{R}_T$ is the transmit vector channel covariance matrix $E\left\{\mathbf{a}_T(t)\mathbf{a}_T^H(t)\right\}P_s$, and P_s is the transmitted mobile power. The well known solution to (2) is the principal eigenvector for $\mathbf{R}_T$ scaled by $\sqrt{P_{T_{max}}}$.

2.3. Invariant Channel Properties

The vector signal received by the base station array due to the mobile user may be written

$$\begin{aligned} \mathbf{x}_R(t) &= \sum_{l=1}^{L}\mathbf{a}_R(\theta_l)\alpha_{R,l}(t)s_R(t)+\mathbf{n}_R(t) \\ &= \mathbf{a}_R(t)s_R(t)+\mathbf{n}_R(t) \end{aligned} \quad (3)$$

The vector $\mathbf{x}_R(t)$ is the base station antenna array output due to the transmitted signal $s_R(t)$ from the mobile user. The additive base station receiver noise vector is $\mathbf{n}_R(t)$.

We wish to determine if replacing $\mathbf{R}_T$ with $\mathbf{R}_R$ in equation (2) will yield a value for received power which is close to the optimum value achieved with known $\mathbf{R}_T$. From Equations (1) and (2), we write

$$\begin{aligned} \mathbf{R}_T &= E\left(\left[\sum_{l=1}^{L}\mathbf{a}_T(\theta_l)\alpha_{T,l}(t)\right]\left[\sum_{k=1}^{L}\mathbf{a}_T(\theta_k)\alpha_{T,k}(t)\right]^H\right) \\ &= \sum_{l=1}^{L}E\left(\left|\alpha_{T,l}(t)\right|^2\right)\mathbf{a}_T(\theta_l)\mathbf{a}_T^H(\theta_l) \end{aligned} \quad (4)$$

where we have assumed that each angle path fades independently so that

$$E\left(\alpha_{T,l}(t)\alpha_{T,k}^*(t)\right)=0 \text{ for } l\neq k \quad (5)$$

Similarly we can write the receive covariance matrix as

$$\mathbf{R}_R=\sum_{l=1}^{L}E\left(\left|\alpha_{R,l}(t)\right|^2\right)\mathbf{a}_R(\theta_l)\mathbf{a}_R^H(\theta_l)+\sigma_n^2\mathbf{I} \quad (6)$$

where σ_n^2 is the receiver noise variance. In equations (4) and (6), we have made the assumptions that the number and angular location of dominant reflectors are the same for transmit and receive frequencies. This is valid since the dominant reflectors are the large features in the radio terrain which tend to dominate the propagation channel and these features are not frequency dependent.

There are three factors which can potentially cause the transmit and receive channel covariance matrices to differ. First, the noise covariance term in (6) is a source for error. With the assumption that the communication environment is capacity limited, this term is insignificant. The second concern is that the average path strengths, $E\left\{|\alpha_l(t)|^2\right\}$, may not be equal. The third source of error is that the array vectors $\mathbf{a}_R(\theta_l)$ and $\mathbf{a}_T(\theta_l)$ can be different.

For most wireless channels, the complex path strength can be modeled as a random process of the form

$$\alpha_{T,l}(t) = \frac{K_T}{d^{\frac{\eta}{2}}}\beta_{R,l}(t)\cdot\sqrt{\Gamma_l}(t) \quad (7)$$

$$\alpha_{R,l}(t) = \frac{K_R}{d^{\frac{\eta}{2}}}\beta_{T,l}(t)\cdot\sqrt{\Gamma_l}(t) \quad (8)$$

where K_T and K_R are the transmit and receive propagation constants which include the effects of antenna gain, transmit power, and carrier wavelength. The distance to the mobile is d. Propagation power loss is proportional to d^η

where η is an empirically determined constant. The effects of slow fading or shadowing are described by the random process $\Gamma_l(t)$ which is not a function of frequency or path angle. This is because the value of the slow fading variable arises directly from the physical properties of the terrain which are not functions of frequency. Fast fading effects are accounted for by the random process $\beta_l(t)$. Slow fading and fast fading are independent random processes. The slow fading variable changes on a much slower scale than the fast fading variable. This allows us to assume that the slow fading is quasi-time invariant over short periods of time. Under these conditions we can write the expected value of the average transmit and receive path path loss conditioned on the distance and slow fading variables as

$$E\left\{\left|\alpha_{T,l}(t)\right|^2/d,\Gamma_l\right\} = \frac{K_T^2}{d^\eta}\Gamma_l E\left\{\left|\beta_{T,l}(t)\right|^2\right\} \quad (9)$$

$$E\left\{\left|\alpha_{R,l}(t)\right|^2/d,\Gamma_l\right\} = \frac{K_T^2}{d^\eta}\Gamma_l E\left\{\left|\beta_{R,l}(t)\right|^2\right\} \quad (10)$$

For most radio channel models, the average strength of the fast fading variable is also insensitive to small changes in frequency. For example, if we use the well known Rayleigh or Rician fading models for $\beta_l(t)$, then $E\left\{|\beta_l(t)|^2\right\}$ is frequency invariant. Without loss of generality, we let $K_T = K_R$ to simplify the expressions that follow. Then we have

$$E\left\{\left|\alpha_{T,l}\right|^2\right\} = E\left\{\left|\alpha_{R,l}\right|^2\right\} = E\left\{|\alpha_l|^2\right\} \quad (11)$$

Equations (4) and (6) express the transmit and receive channel covariance as a weighted summation of rank 1 positive semidefinite matrices of the form $\mathbf{a}(\theta_l)\mathbf{a}^H(\theta_l)$. We have established that the weights $(E[|\alpha_l|^2])$ and the path angles θ_l are invariant to the duplex frequency translation. If we can show that the array response vector is also invariant, then we can use $\mathbf{R}_R$ in place of $\mathbf{R}_T$ in (2) with confidence. For an arbitrary selection of array geometries, the transmit and receive array responses vectors can differ substantially. However, there are two methods for minimizing the difference in array response:

- Design two separate closely located arrays which each have similar response. We term this the matched array approach.
- Design a single array which minimizes the difference between the transmit response and the receive response. We term this the duplex array approach.

An example of the matched array approach is for the transmit array to be a scaled version of the receive array, the scale factor being the ratio of transmit to receive wavelength. In practice it is not possible to manufacture the two arrays to be perfectly matched. Also, in some environments, there are physical objects located in the near field of the array. This causes the transmit and receive angle spectra to differ because the arrays probe different locations in the scattering field. This becomes less important if the antenna set is located high above the terrain with no near-field obstructions.

The potential for imperfect array matching and near-field effects motivate us to consider the duplex array approach. We investigate the difference between transmit and receive array response vectors assuming omni azimuth directional radiating elements. The general expression for the receive array response vector, $\mathbf{a}_R(\theta)$, with omni directional elements is

$$\mathbf{a}_R(\theta) = \begin{bmatrix}1 & e^{j\psi_{R,2}} \dots & e^{j\psi_{R,M}}\end{bmatrix}^T \quad (12)$$

$$\psi_{R,i} = \frac{2\pi}{\lambda_R} d_{1,i} sin(\theta_{1,i}) \quad (13)$$

where $\theta_{1,i}$ is the path angle with respect to the perpendicular to the line joining element 1 and element i, $d_{1,i}$ is the distance between element 1 and element i, and λ_R is the receive carrier wavelength. For *small frequency shifts*, we can write the transmit array response vector as a linear perturbation of the receive vector as follows.

$$\begin{aligned}\mathbf{a}_T(\theta) &= \begin{bmatrix}1 & e^{j\psi_{T,2}} \dots & e^{j\psi_{T,M}}\end{bmatrix}^T \quad (14)\\ &= \mathbf{\Phi}\mathbf{a}_R(\theta)\\ \mathbf{\Phi} &= diag(1, e^{j\phi_2}, \dots, e^{j\phi_M})\\ \phi_i &= \left(\frac{1}{\lambda_T} - \frac{1}{\lambda_R}\right)\psi_{R,i}\end{aligned}$$

Substituting (11) and (14) into (4), we express the transmit covariance matrix as

$$\mathbf{R}_T = \sum_{l=1}^{L} \mathbf{E}\left(|\alpha_l(t)|^2\right)\mathbf{\Phi}_l\mathbf{a}_R(\theta_l)\mathbf{a}_R^H(\theta_l)\mathbf{\Phi}_l^H \quad (15)$$

where each rotation matrix $\mathbf{\Phi}_l$ corresponds to the angle θ_l. Thus, for the omni-directional element duplex array, the difference between transmit and receive covariance matrices is due to small rotational perturbations on each of the component array response vectors. We note that if the complex M-space rotation angles due to the $\mathbf{\Phi}_l$ are bounded to small values, we do not expect the eigenvalues or the eigenvectors of the two matrices to be drastically different (*i.e.* the small rotations will not have a large impact on the subspace occupied by the channel vectors.

We make the approximation that the path angle spread is confined so that we can write

$$\mathbf{\Phi}_l \approx \mathbf{\Phi}_0 \quad \forall l \quad (16)$$

resulting in the received mobile power expression

$$P_{rcv} \approx \mathbf{w}^H\mathbf{\Phi}_0\mathbf{R}_R\mathbf{\Phi}_0^H\mathbf{w} \quad (17)$$

We know that if we substitute $\mathbf{R}_R$ for $\mathbf{R}_T$ and solve (2), the solution will be

$$\mathbf{w_T} = \mathbf{v}_1(\mathbf{R}_R)\sqrt{P_{Tmax}} \quad (18)$$

where $\mathbf{v}_1(\mathbf{R}_R)$ is the principal eigenvector of $\mathbf{R}_R$. Using (17) and (16),

$$P_{rcv} \approx P_{T_{max}}\mathbf{v}_1^H\mathbf{\Phi}_0\sum_{i=1}^{M}\sigma_i^2\mathbf{v}_i\mathbf{v}_i^H\mathbf{\Phi}_0^H\mathbf{v}_1 \quad (19)$$

where σ_i^2 and $\mathbf{v}_i$ are the i^{th} eigenvalue and eigenvector of $\mathbf{R}_R$ respectively. Defining ϕ_{max} as the maximum complex phase shift between any two diagonal elements of $\mathbf{\Phi}_0$ over all possible central angles of arrival (θ_0) for the given array geometry, we can easily show that (19) leads to

$$P_{rcv} \gtrsim P_{Tmax}\sigma_1^2 cos(\phi_{max}) \qquad (20)$$

for $\phi_{max} \leq \frac{\pi}{2}$. For an arbitrary antenna array which has unlimited maximum element spacing, the approximate bound given by (20) can be arbitrarily small. However, if we choose an appropriate array geometry, the bound is useful. For example, with an eight element circular array with a minimum inter-element spacing of $\frac{\lambda}{2}$ and a receive to transmit frequency shift of 6%, the bound yields a maximum received power degradation of approximately 0.5dB when compared to the power that would be delivered if $\mathbf{R}_R$ were equal to $\mathbf{R}_T$.

We conclude that if we carefully design the antenna system, a near-optimal solution can be found for the single antenna transmit weight vector problem using a covariance characterization of the received channel vector. The limit on maximum element spacing inferred by (20) constrains the number of array elements and the array gain. While we do not discuss it here, it is possible to relax the constraint on element spacing (and array gain) by employing directive elements patterns so that elements which are far apart do not radiate to the same portion of the angular spectrum. An example of this would be a circular array with cardioid element patterns which overlap.

3. COOPERATIVE TRANSMISSION NETWORK

We now address the problem of improving spectral efficiency (Erlangs/Hz/Km^2) of a wireless network by forming transmit and receive beam patterns which maximize desired user power while minimizing undesired user interference power. We choose to improve the frequency re-use separation required between base stations rather than to share spectrum between multiple co-channel users within the same cell as proposed in [2, 1]. We term the same-cell co-channel re-use problem spatial division multiple access (SDMA) and we term our re-use improvement approach the cooperative transmission network (CTN).

The CTN approach has several advantages. We only touch on the most significant issues here. In SDMA, it is necessary to estimate the transmit channel properties of all co-channel users. This is extremely difficult considering the large dynamic range that wireless channels operate over (>70dB). In SDMA, we must also construct a base station receive beam which will null the strongest signal so that the weakest can be received. This requires null depths which are impractical. In CTN, each base station need only explicitly estimate one desired user. Undesired interference channel behavior may can be found without explicitly estimating each undesired channel. This is discussed further below. Since co-channel interference in the CTN approach is due to adjacent cells and is never due to a use that is very close to the desired base station, the base station receiver dynamic range problem is avoided. In summary, the SDMA approach suffers due to the ill behaved nature of the multipath propagation environment. In contrast, the CTN approach actually takes advantage of the essential nature of the environment to make the beam pattern optimization problem easier.

3.1. Cooperative Algorithm

We formulate the re-use improvement problem into an optimization that is relatively inexpensive to solve and eliminates the need for knowledge of the individual adjacent user covariance matrices. Assuming (without loss of generality) that the radiated power for each mobile user is 1, we write the total received data vector covariance matrix as the sum of the desired user's covariance, the undesired user's covariance, and the noise covariance

$$\mathbf{R}_x = E\left\{\mathbf{x}_R(t)\mathbf{x}_R^H(t)\right\} = \mathbf{R}_{R,d} + \sum_{\substack{j=1 \\ j\neq d}}^{U} \mathbf{R}_{R,j} + \mathbf{R}_n \qquad (21)$$

We can then estimate the undesired radiated interference covariance with

$$\hat{\mathbf{R}}_{I,R} = \hat{\mathbf{R}}_x - \hat{\mathbf{R}}_{R,d} \qquad (22)$$

where $\hat{\mathbf{R}}_x$, $\hat{\mathbf{R}}_{R,d}$ and $\hat{\mathbf{R}}_{I,R}$ are the estimated covariance matrices. We have again assumed that the cellular network is operating in a capacity limited environment. Equation (22) leads to an optimization which depends on the summation of the interference covariance matrices rather than the individual interference covariance matrices:

$$\hat{\mathbf{w}} = \arg\left\{\max_{\mathbf{w}}\left(\mathbf{w}^H\mathbf{R}_{T,d}\mathbf{w}\right)\right\} \qquad (23)$$

s.t.

$$\mathbf{w}^H\mathbf{R}_{I,T}\mathbf{w} =$$

$$\mathbf{w}^H\left[\left(\sum_{\substack{j=1 \\ i\neq d}}^{U}\mathbf{R}_{T,j}\right) + \mathbf{I}_M\frac{P_{Imax}}{P_{Tmax}}\right]\mathbf{w} \leq P_{Imax}$$

followed by the feasibility checks

$$\mathbf{w}^H\mathbf{w} \geq P_{Tmin}$$

$$\mathbf{w}^H\mathbf{R}_{T,d}\mathbf{w} \geq P_{Rmin}$$

Note that in (23), the maximum transmitted power constraint has been blended with the summation of the maximum interference constraints. The solution to (23) is the scaled generalized eigenvector associated with the largest generalized eigenvalue of the matrix pair $[\mathbf{R}_{T,d}, \mathbf{R}_{I,T}]$.

3.2. Substituting Receive Covariance

We now consider the effects of substituting receive covariance for transmit covariance in the multiple user transmitter problem. The effects of imperfect transmit covariance estimates will have a more severe impact on multiple user radiated interference performance than on single user received power performance. In the single user case the effect of imperfect covariance estimates is a slight decrease in desired user power. This is because the small perturbations cause the dominant eigenvectors of $\mathbf{R}_T$ to be slightly perturbed versions of the dominant eigenvectors of $\mathbf{R}_R$. Hence, the inner product between the dominant eigenvector of $\mathbf{R}_R$

and $\mathbf{R}_T$ is still large. In contrast, the solution to (23) yields a weight vector which has as high an inner product as possible on $\mathbf{R}_d$ while maintaining a *very small inner product* on $\mathbf{R}_I$. In this case, the slight perturbations of the dominant eigenvectors of the interference covariance matrix $\mathbf{R}_I$ can cause a large relative increase in the interference power.

Thus, we expect that the signal to interference ratio achieved at a given mobile will be degraded from the minimum value set by the constraints P_{Imax} and P_{Rmin} due to differences in the array response. We observe that the degradation will be dependent on several factors such as desired user position, undesired user position, and angle of arrival spread. These parameters are random variables so we conclude that the degradation will be a random variable with some associated distribution. Therefore, we pursue the strategy of increasing the constraint conditions by an amount sufficient to offset the statistical degradation caused by transmit covariance matrix estimation errors. We evaluate the effectiveness of this approach through a set of extensive simulations in Section 4. It should be noted that it is possible to adjust network spectrum capacity Vs. quality trade-offs simply by varying the constraint values P_{Rmin} and P_{Imax} in (23).

4. CELLULAR CAPACITY IMPROVEMENT SIMULATION RESULTS

The simulation compared the SINR performance of three transmit antenna scenarios:

- The matched array approach with no matching errors.
- The duplex array approach including frequency translation errors.
- A single omni-directional antenna.

The simulation includes the effects of adjacent cell users, distance loss, shadow fading, fast fading, and changes in mobile position. A re-use factor of 1 was chosen for this simulation. That is, each frequency channel is re-used at each base station site with no separation between co-channel cells. Many other re-use factors are possible. Unity re-use was chosen because it is the most severe test case.

The simulated array geometry is an 8 element circular array with minimum element spacing of $\frac{\lambda_R}{2}$. The array vectors are computed based on omni directional elements and the transmit wavelength is reduced 6% from the receive wavelength. The decisions made at one base station in the network impact the SINR performance of the mobiles in adjoining cells. Therefore, it is necessary to include several base stations in the simulation. For a frequency reuse interval of 1, it is necessary to include at least 19 cell sites in the simulation to account for the effects of first tier interferers. The base stations are distributed as shown in Figure 2.

For each experiment, the 19 users are located within once cell radius of their desired base station according to an independent uniform spatial distribution. The central angle for the vector channel from each base to each user is determined by the relative location of the user and base. There are 10 independently fading paths from each base station to each user, uniformly distributed around the central angle according to the angle spread parameter. Each of the 10 paths between each user and each base has iid Rayleigh fading. Each path also has log-normal fading which is iid over different base station to mobile user channels, but is partially correlated between the 10 paths from a given mobile to a given base station. The marginal log-normal standard deviation is 8dB. All paths are also weighted with the distance loss using an exponent of 4.

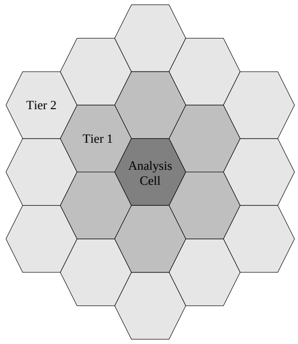


Figure 2: Cellular Simulation Diagram. The Analysis Cell base station forms a beam with the Analysis Cell user as the objective and Tier 1 cell users as constraints. The Tier 1 cell base stations form beams with their desired user as the objective and adjacent users from Analysis Cell, Tier 1 cells, and Tier 2 cells as constraints. No base stations are present in Tier 2– only users which form constraints for Tier 1 cells.

Samples of the transmit and receive array vectors for a given propagation link between mobile i and base j are generated by

$$\hat{\mathbf{a}}_{R,i,j}(kT) = \frac{K_R}{d_{i,j}^{\frac{\eta}{2}}}\left[\sum_{l=1}^{L}\beta_{R,i,j,l}(kT)\mathbf{a}_R(\theta_{i,j,l})\sqrt{\Gamma_{i,j,l}}\right] \quad (24)$$

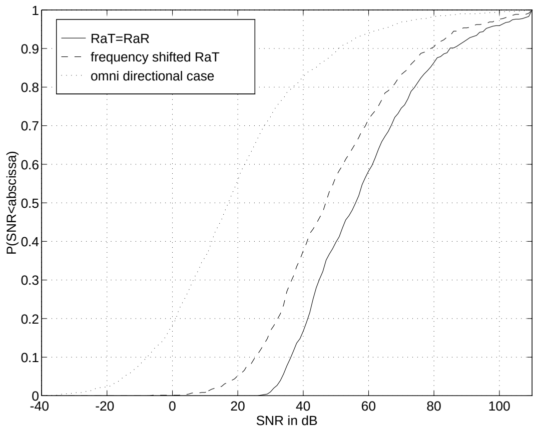
$$\hat{\mathbf{a}}_{T,i,j}(kT) = \frac{K_T}{d_{i,j}^{\frac{\eta}{2}}}\left[\sum_{l=1}^{L}\beta_{T,i,j,l}(kT)\mathbf{a}_T(\theta_{i,j,l})\sqrt{\Gamma_{i,j,l}}\right] \quad (25)$$

The array vector snapshots are used to estimate the receive covariance matrix for each user i :

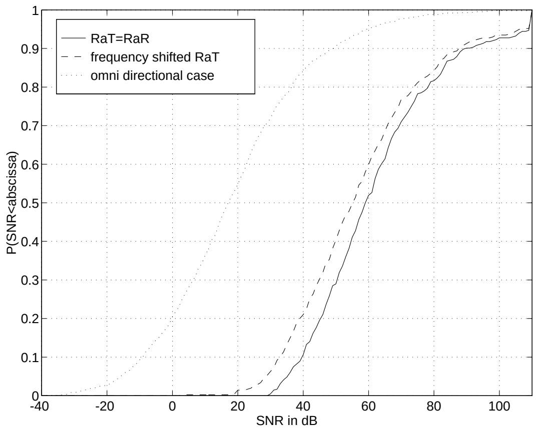
$$\hat{\mathbf{R}}_{R,i,j}(KT) = \frac{1}{2N_w}\sum_{k=K-N_w}^{K+N_w-1}\hat{\mathbf{a}}_{i,j}(kT)\hat{\mathbf{a}}_{i,j}^H(kT) \quad (26)$$

Ten array vector snapshots are used to estimate each covariance matrix. Two weight vector solutions to Equation (23) are computed. One solution is computed for the perfectly matched array. Another solution is computed for the duplex array with the substitution $\hat{\mathbf{R}}_{R,i}(KT) \rightarrow \hat{\mathbf{R}}_{T,i}(KT)$. Of the 19 cells considered in the analysis, the full SINR statistics are collected only for the central cell user.

The important quantities for each simulation run are the desired power delivered by the central cell base station to the central cell user, and the interference power delivered by each of the first tier base stations to the central cell user. If the feasibility checks, based on receive covariance, are not satisfied, then the algorithm concludes that the present frequency channel is unsuitable for co-channel sharing for the user under analysis. In this way, each base station only chooses to use a channel that is populated by adjacent mobiles possessing interference channels that are compatible with the desired user. If the checks are not satisfied at the central cell base station, an unsuccessful co-channel transmission attempt is logged. This is a pessimistic assumption since a frequency allocation algorithm could find another user that is compatible with the unused channel. Statistics are collected for SINR cumulative distribution function (cdf) and percentage of frequency channel sharing successes over 1,000 experiments.



(a)



(b)

Figure 3: (a) SINR cdf for angle spread of 10°, success rate=71%. (b) SINR cdf for angle spread of 30°, success rate=46%.

The simulation results are presented for each of two different angle spread conditions in Figure 3. Figure 3(a) displays the cdf data for an angle spread of 10° for each user. Figure 3(b) displays the cdf data for an angle spread of 30°. The frequency sharing success rate was 71% for the 10° angle spread case and 46% for the 30° angle spread case. This equates to corresponding frequency re-use factors of 1.4 and 2.2 respectively.

The results of the simulation indicate an improvement of better than 50dB for the perfectly matched matched array and 30dB for the duplex array as compared to the omni directional case at the 2% probability of outage point in the cdfs. While it is impractical to obtain perfect matching for a realistic matched array, the matched array results provide insight into the degradation resulting from frequency translation in the duplex array approach.

Presently, most cellular networks use a frequency re-use factor of 7. The results presented here suggest that a frequency spectrum capacity improvement factor in the range of 3 to 6 may be possible using our transmit beam forming approach. As mentioned above, these simulations did not account for additional capacity improvement that would occur if a frequency allocation algorithm were implemented. Thus the capacity improvement factors may be quite conservative.

5. CONCLUSION

Forming an optimum adaptive transmission beam pattern in the presence of multipath propagation has been a longstanding and difficult problem. In this paper, we have proposed a method for solving this problem without any explicit knowledge of the array geometry or mobile feedback. We utilize receive vector channel estimates to form the transmission beam pattern. We showed through analysis that using this strategy works well in the single-user case. We also demonstrated through simulations that our cooperative transmission network multiple user approach can potentially increases re-use capacity by a factor of 3 to 6. These preliminary results encourage us to believe that this approach holds substantial promise. The basic transmission beam forming techniques presented here have been extended to the delay-spread digital channel and the CDMA channel. These new results will be reported in a future paper.

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