

[00111] U.S. Patent No. 7,418,053, filed July 30, 2004, issued August 26, 2008, entitled "System and Method for Distributed Input Distributed Output Wireless Communication";

[00112] U.S. Application Serial No. 10/817,731, filed April 2, 2004 entitled "System and Method For Enhancing Near Vertical Incidence Skywave ("NVIS") Communication Using Space-Time Coding.

[00113] To reduce the size and complexity of the present patent application, the disclosure of some of the related patents and applications is not explicitly set forth below. Please see the related patents and applications for a full detailed description of the disclosure.

[00114] Note that section I below (Disclosure From Related Application Serial No. 12/802,988) utilizes its own set of endnotes which refer to prior art references and prior applications assigned to the assignee of the present application. The endnote citations are listed at the end of section I (just prior to the heading for Section II). Citations in Section II uses may have numerical designations for its citations which overlap with those used in Section I even through these numerical designations identify different references (listed at the end of Section II). Thus, references identified by a particular numerical designation may be identified within the section in which the numerical designation is used.

I. Disclosure From Related Application Serial No. 12/802,988

1. Methods to Remove Inter-cluster Interference

Described below are wireless radio frequency (RF) communication systems and methods employing a plurality of distributed transmitting antennas to create locations in space with zero RF energy. When M transmit antennas are employed, it is possible to create up to (M-1) points of zero RF energy in predefined locations. In one embodiment of the invention, the points of zero RF energy are wireless devices and the transmit antennas are aware of the channel state information (CSI) between the transmitters and the receivers. In one embodiment, the CSI is computed at the receivers and fed back to the transmitters. In another embodiment, the CSI is computed at the transmitter via training from the receivers, assuming channel reciprocity is exploited. The transmitters may utilize the CSI to determine the interfering signals to be simultaneously transmitted. In one embodiment, block diagonalization (BD) precoding is employed at the transmit antennas to generate points of zero RF energy.

[00115] The system and methods described herein differ from the conventional receive/transmit beamforming techniques described above. In fact, receive beamforming computes the weights to suppress interference at the receive side (via null-steering), whereas some embodiments of the invention described herein apply weights at the transmit side to create interference patterns that result in one or multiple locations in space with "zero RF energy." Unlike conventional transmit beamforming or BD precoding designed to maximize

the system increases linearly with the number of subcarriers. For example, in OFDM systems with 1MHz signal bandwidth, the cyclic prefix (L_0) must have at least eight channel taps (i.e., duration of 8 microseconds) to avoid intersymbol interference in outdoor urban macrocell environments with large delay spread [3]. The size (N_{FFT}) of the fast Fourier transform (FFT) used to generate the OFDM symbols is typically set to multiple of L_0 to reduce loss of data rate. If $N_{\text{FFT}}=64$, the effective spectral efficiency of the system is limited by a factor $N_{\text{FFT}}/(N_{\text{FFT}}+L_0)=89\%$. Larger values of N_{FFT} yield higher spectral efficiency at the expense of higher computational complexity at the DIDO precoder.

[00194] One way to reduce computational complexity at the DIDO precoder is to carry out the SVD operation over a subset of tones (that we call *pilot tones*) and derive the precoding weights for the remaining tones via interpolation. Weight interpolation is one source of error that results in inter-client interference. In one embodiment, optimal weight interpolation techniques are employed to reduce inter-client interference, yielding improved error rate performance and lower computational complexity in multicarrier systems. In DIDO systems with M transmit antennas, U clients and N receive antennas per clients, the condition for the precoding weights of the k^{th} client ($\mathbf{W}_k$) that guarantees zero interference to the other clients u is derived from (2) as

$$\mathbf{H}_u \mathbf{W}_k = \mathbf{0}^{N \times N}; \quad \forall u = 1, \dots, U; \text{ with } u \neq k \quad (24)$$

where $\mathbf{H}_u$ are the channel matrices corresponding to the other DIDO clients in the system.

[00195] In one embodiment of the invention, the objective function of the weight interpolation method is defined as

$$f(\boldsymbol{\theta}_k) = \sum_{\substack{u=1 \\ u \neq k}}^U \|\mathbf{H}_u \hat{\mathbf{W}}_k(\boldsymbol{\theta}_k)\|_F \quad (25)$$

where $\boldsymbol{\theta}_k$ is the set of parameters to be optimized for user k , $\hat{\mathbf{W}}_k(\boldsymbol{\theta}_k)$ is the *weight interpolation matrix* and $\|\cdot\|_F$ denotes the Frobenius norm of a matrix. The optimization problem is formulated as

$$\boldsymbol{\theta}_{k,\text{opt}} = \arg \min_{\boldsymbol{\theta}_k \in \Theta_k} f(\boldsymbol{\theta}_k) \quad (26)$$

where Θ_k is the feasible set of the optimization problem and $\boldsymbol{\theta}_{k,\text{opt}}$ is the optimal solution.

[00196] The objective function in (25) is defined for one OFDM tone. In another embodiment of the invention, the objective function is defined as linear combination of the Frobenius norm in (25) of the matrices for all the OFDM tones to be interpolated. In another embodiment, the OFDM spectrum is divided into subsets of tones and the optimal solution is given by

$$\boldsymbol{\theta}_{k,\text{opt}} = \arg \min_{\boldsymbol{\theta}_k \in \Theta_k} \max_{n \in A} f(n, \boldsymbol{\theta}_k) \quad (27)$$

where n is the OFDM tone index and A is the subset of tones.

[00197] The weight interpolation matrix $\mathbf{W}_k(\boldsymbol{\theta}_k)$ in (25) is expressed as a function of a set of parameters $\boldsymbol{\theta}_k$. Once the optimal set is determined according to (26) or (27), the optimal weight matrix is computed. In one embodiment of the invention, the weight interpolation matrix of given OFDM tone n is defined as linear combination of the weight matrices of the pilot tones. One example of weight interpolation function for beamforming systems with single client was defined in [11]. In DIDO multi-client systems we write the weight interpolation matrix as

$$\hat{\mathbf{W}}_k(lN_0 + n, \theta_k) = (1 - c_n) \cdot \mathbf{W}(l) + c_n e^{j\theta_k} \cdot \mathbf{W}(l+1) \quad (28)$$

where $0 \leq l \leq (L_0-1)$, L_0 is the number of pilot tones and $c_n = (n-1)/N_0$, with $N_0 = N_{FFT}/L_0$. The weight matrix in (28) is then normalized such that $\|\hat{\mathbf{W}}_k\|_F = \sqrt{NM}$ to guarantee unitary power transmission from every antenna. If $N=1$ (single receive antenna per client), the matrix in (28) becomes a vector that is normalized with respect to its norm. In one embodiment of the invention, the pilot tones are chosen uniformly within the range of the OFDM tones. In another embodiment, the pilot tones are adaptively chosen based on the CSI to minimize the interpolation error.

[00198] We observe that one key difference of the system and method in [11] against the one proposed in this patent application is the objective function. In particular, the systems in [11] assumes multiple transmit antennas and single client, so the related method is designed to maximize the product of the precoding weight by the channel to maximize the receive SNR for the client. This method, however, does not work in multi-client scenarios, since it yields inter-client interference due to interpolation error. By contrast, our method is designed to minimize inter-client interference thereby improving error rate performance to all clients.

c. **CSI report to the BTSs:** Once the user clusters are selected, the CSI from all transmitters within the user-cluster to every user reached by those transmitters is made available to all BTSs. The CSI information is shared across all BTSs via the BSN. In TDD systems, UL/DL channel reciprocity can be exploited to derive the CSI from training over the UL channel. In FDD systems, feedback channels from all users to the BTSs are required. To reduce the amount of feedback, only the CSI corresponding to the non-zero entries of the link-quality matrix are fed back.

d. **DIDO precoding:** Finally, DIDO precoding is applied to every CSI sub-matrix corresponding to different user clusters (as described, for example, in the related U.S. Patent Applications).

In one embodiment, singular value decomposition (SVD) of the effective channel matrix $\tilde{\mathbf{H}}_k$ is computed and the precoding weight $\mathbf{w}_k$ for user k is defined as the right singular vector corresponding to the null subspace of $\tilde{\mathbf{H}}_k$. Alternatively, if $M > K$ and the SVD decomposes the effective channel matrix as $\tilde{\mathbf{H}}_k = \mathbf{V}_k \mathbf{\Sigma}_k \mathbf{U}_k^H$, the DIDO precoding weight for user k is given by

$$\mathbf{w}_k = \mathbf{U}_0 (\mathbf{U}_0^H \cdot \mathbf{h}_k^T)$$

where $\mathbf{U}_0$ is the matrix with columns being the singular vectors of the null subspace of $\tilde{\mathbf{H}}_k$.

From basic linear algebra considerations, we observe that the right singular vector in the null subspace of the matrix $\tilde{\mathbf{H}}$ is equal to the eigenvector of $\mathbf{C}$ corresponding to the zero eigenvalue

$$\mathbf{C} = \tilde{\mathbf{H}}^H \tilde{\mathbf{H}} = (\mathbf{V} \mathbf{\Sigma} \mathbf{U}^H)^H (\mathbf{V} \mathbf{\Sigma} \mathbf{U}^H) = \mathbf{U} \mathbf{\Sigma}^2 \mathbf{U}^H$$

where the effective channel matrix is decomposed as $\tilde{\mathbf{H}} = \mathbf{V} \mathbf{\Sigma} \mathbf{U}^H$, according to the SVD. Then, one alternative to computing the SVD of $\tilde{\mathbf{H}}_k$ is to calculate the eigenvalue decomposition of $\mathbf{C}$. There are several methods to compute eigenvalue decomposition such as the power method. Since we are only interested to the eigenvector corresponding to the null subspace of $\mathbf{C}$, we use the inverse power method described by the iteration

$$\mathbf{u}_{i+1} = \frac{(\mathbf{C} - \lambda \mathbf{I})^{-1} \mathbf{u}_i}{\|(\mathbf{C} - \lambda \mathbf{I})^{-1} \mathbf{u}_i\|}$$

where the vector $(\mathbf{u}_i)$ at the first iteration is a random vector.

Given that the eigenvalue (λ) of the null subspace is known (i.e., zero) the inverse power method requires only one iteration to converge, thereby reducing computational complexity. Then, we write the precoding weight vector as

$$\mathbf{w} = \mathbf{C}^{-1} \mathbf{u}_1$$

where $\mathbf{u}_1$ is the vector with real entries equal to 1 (i.e., the precoding weight vector is the sum of the columns of $\mathbf{C}^{-1}$).

The DIDO precoding calculation requires one matrix inversion. There are several numerical solutions to reduce the complexity of matrix inversions such as the Strassen's algorithm [1] or the Coppersmith-Winograd's algorithm [2,3]. Since $\mathbf{C}$ is Hermitian matrix by definition, an alternative solution is to decompose $\mathbf{C}$ in its real and imaginary components and compute matrix inversion of a real matrix, according to the method in [4, Section 11.4].

[00238] Another feature of the proposed method and system is its reconfigurability. As the client moves across different DIDO clusters as in **Figure 42**, the user-cluster follows its moves. In other words, the subset of transmit antennas is constantly updated as the client changes its position and the effective channel matrix (and corresponding precoding weights) are recomputed.

[00239] The method proposed herein works within the super-cluster in **Figure 36**, since the links between the BTSs via the BSN must be low-latency. To suppress interference in the overlapping regions of different super-clusters, it is possible to use our method in [5] that uses extra antennas to create points of zero RF energy in the interfering regions between DIDO clusters.

[00240] It should be noted that the terms "user" and "client" are used interchangeably herein.

References

[00241] [1] S. Robinson, "Toward an Optimal Algorithm for Matrix Multiplication", SIAM News, Volume 38, Number 9, November 2005.

[00242] [2] D. Coppersmith and S. Winograd, "Matrix Multiplication via Arithmetic Progression", J. Symb. Comp. vol.9, p.251-280, 1990.

[00243] [3] H. Cohn, R. Kleinberg, B. Szegedy, C. Umans, "Group-theoretic Algorithms for Matrix Multiplication", p. 379-388, Nov. 2005.

[00244] [4] W.H. Press, S.A. Teukolsky, W. T. Vetterling, B.P. Flannery "NUMERICAL RECIPES IN C: THE ART OF SCIENTIFIC COMPUTING", Cambridge University Press, 1992.

[00245] [5] A. Forenza and S.G. Perlman, "INTERFERENCE MANAGEMENT, HANDOFF, POWER CONTROL AND LINK ADAPTATION IN DISTRIBUTED-INPUT DISTRIBUTED-OUTPUT (DIDO) COMMUNICATION SYSTEMS", Patent Application Serial No. 12/802,988, filed June 16, 2010.

[00246] [6] Per-Erik Eriksson and Björn Odenhammar, "VDSL2: Next important broadband technology", Ericsson Review No. 1, 2006.

III. SYSTEMS AND METHODS TO EXPLOIT AREAS OF COHERENCE IN WIRELESS SYSTEMS

[00247] The capacity of multiple antenna systems (MAS) in practical propagation environments is a function of the spatial diversity available over the wireless link. Spatial diversity is determined by the distribution of scattering objects in the wireless channel as well as the geometry of transmit and receive antenna arrays.

[00248] One popular model for MAS channels is the so called clustered channel model, that defines groups of scatterers as clusters located around the transmitters and receivers. In general, the more clusters and the larger their angular spread, the higher spatial diversity and capacity achievable over wireless links. Clustered channel models have been validated through practical measurements [1-2] and variations of those models have been adopted by different indoor (i.e., IEEE 802.11n Technical Group [3] for WLAN) and outdoor (3GPP Technical Specification Group for 3G cellular systems [4]) wireless standards.

[00249] Other factors that determine the spatial diversity in wireless channels are the characteristics of the antenna arrays, including: antenna element spacing [5-7], number of antennas [8-9], array aperture [10-11], array geometry [5, 12, 13], polarization and antenna pattern [14-28].

[00250] A unified model describing the effects of antenna array design as well as the characteristics of the propagation channel on the spatial diversity (or degrees of freedom) of wireless links was presented in [29]. The received signal model in [29] is given by

$$\underline{y}(\underline{q}) = \int \underline{C}(\underline{q}, \underline{p}) \underline{x}(\underline{p}) d\underline{p} + \underline{z}(\underline{q})$$

where $\underline{x}(\underline{p}) \in \mathbb{C}^3$ is the polarized vector describing the transmit signal, $\underline{p}, \underline{q} \in \mathbb{R}^3$ are the polarized vector positions describing the transmit and receive arrays, respectively, and $\underline{C}(\cdot, \cdot) \in \mathbb{C}^{3 \times 3}$ is the matrix describing the system response between transmit and receive vector positions given by

$$\underline{C}(\underline{q}, \underline{p}) = \iint \underline{A}_r(\underline{q}, \underline{\hat{m}}) \underline{H}(\underline{\hat{m}}, \underline{\hat{n}}) \underline{A}_t(\underline{\hat{n}}, \underline{p}) d\underline{\hat{n}} d\underline{\hat{m}}$$

where $\underline{A}_t(\cdot, \cdot), \underline{A}_r(\cdot, \cdot) \in \mathbb{C}^{3 \times 3}$ are the transmit and receive array responses respectively and $\underline{H}(\underline{\hat{m}}, \underline{\hat{n}}) \in \mathbb{C}^{3 \times 3}$ is the channel response matrix with entries being the complex gains between transmit direction $\underline{\hat{n}}$ and receive direction $\underline{\hat{m}}$. In DIDO systems, user devices may have single or multiple antennas. For the sake of simplicity, we assume single antenna receivers with ideal isotropic patterns and rewrite the system response matrix as

$$\underline{\mathbf{C}}(\underline{\mathbf{q}}, \underline{\mathbf{p}}) = \int \underline{\mathbf{H}}(\underline{\mathbf{q}}, \underline{\mathbf{n}}) \underline{\mathbf{A}}(\underline{\mathbf{n}}, \underline{\mathbf{p}}) d\underline{\mathbf{n}}$$

where only the transmit antenna pattern $\underline{\mathbf{A}}(\underline{\mathbf{n}}, \underline{\mathbf{p}})$ is considered.

[00251] From the Maxwell equations and the far-field term of the Green function, the array response can be approximated as [29]

$$\underline{\mathbf{A}}(\underline{\mathbf{n}}, \underline{\mathbf{p}}) = \frac{j\eta e^{j2\pi d_0}}{2\lambda^2 d_0} (\underline{\mathbf{I}} - \underline{\mathbf{n}}\underline{\mathbf{n}}^H) \underline{\mathbf{a}}(\underline{\mathbf{n}}, \underline{\mathbf{p}})$$

with $\underline{\mathbf{p}} \in P$, P is the space that defines the antenna array and where

$$\underline{\mathbf{a}}(\underline{\mathbf{n}}, \underline{\mathbf{p}}) = \exp(-j2\pi \underline{\mathbf{n}}^H \underline{\mathbf{p}})$$

with $(\underline{\mathbf{n}}, \underline{\mathbf{p}}) \in \Omega \times P$. For unpolarized antennas, studying the array response is equivalent to study the integral kernel above. Hereafter, we show closed for expressions of the integral kernels for different types of arrays.

Unpolarized Linear Arrays

[00252] For unpolarized linear arrays of length L (normalized by the wavelength) and antenna elements oriented along the z -axis and centered at the origin, the integral kernel is given by [29]

$$\underline{\mathbf{a}}(\cos \underline{\theta}, \underline{p}_z) = \exp(-j2\pi \underline{p}_z \cos \underline{\theta}).$$

[00253] Expanding the above equation into a series of shifted dyads, we obtain that the sinc function have resolution of $1/L$ and the dimension of the array-limited and approximately wavevector-limited subspace (i.e., degrees of freedom) is

$$D_F = L |\Omega_\theta|$$

where $\Omega_\theta = \{\cos \theta : \theta \in \Theta\}$. We observe that for broadside arrays $|\Omega_\theta| = |\Theta|$ whereas for endfire $|\Omega_\theta| \approx |\Theta|^2/2$.

Unpolarized Spherical Arrays

[00254] The integral kernel for a spherical array of radius R (normalized by the wavelength) is given by [29]

$$\underline{\mathbf{a}}(\underline{\mathbf{n}}, \underline{\mathbf{p}}) = \exp\{-j2\pi R [\sin \underline{\theta} \sin \underline{\theta}' \cos(\underline{\phi} - \underline{\phi}') + \cos \underline{\theta} \cos \underline{\theta}']\}.$$

[00255] Decomposing the above function with sum of spherical Bessel functions of the first kind we obtain the resolution of spherical arrays is $1/(\pi R^2)$ and

the degrees of freedom are given by

$$D_F = \underline{A}|\Omega| = \pi R^2 |\Omega|$$

[00256] where A is the area of the spherical array and $|\Omega| \subset [0, \pi) \times [0, 2\pi)$.

Areas of Coherence in Wireless Channels

[00257] The relation between the resolution of spherical arrays and their area A is depicted in **Figure 43**. The sphere in the middle is the spherical array of area A . The projection of the channel clusters on the unit sphere defines different scattering regions of size proportional to the angular spread of the clusters. The area of size $1/A$ within each cluster, which we call “area of coherence”, denotes the projection of the basis functions of the radiated field of the array and defines the resolution of the array in the wavevector domain.

[00258] Comparing **Figure 43** with **Figure 44**, we observe that the size of the area of coherence decreases as the inverse of the size of the array. In fact, larger arrays can focus energy into smaller areas, yielding larger number of degrees of freedom D_F . Note that total number of degrees of freedom depends also on the angular spread of the cluster, as shown in the definition above.

[00259] **Figure 45** depicts another example where the array size covers even larger area than **Figure 44**, yielding additional degrees of freedom. In DIDO systems, the array aperture can be approximated by the total area covered by all DIDO transmitters (assuming antennas are spaced fractions of wavelength apart). Then **Figure 45** shows that DIDO systems can achieve increasing numbers of degrees of freedom by distributing antennas in space, thereby reducing the size of the areas of coherence. Note that these figures are generated assuming ideal spherical arrays. In practical scenarios, DIDO antennas spread random across wide areas and the resulting shape of the areas of coherence may not be as regular as in the figures.