Increasing Wireless Channel Capacity Through MIMO Systems Employing Co-Located Antennas

Anand S. Konanur, *Student Member, IEEE*, Keyoor Gosalia, *Member, IEEE*, Sandeep H. Krishnamurthy, *Student Member, IEEE*, Brian Hughes, *Member, IEEE*, and Gianluca Lazzi, *Senior Member, IEEE*

***Abstract*—Wireless networks consisting of compact antennas find applications in diverse areas such as communication systems, direction of arrival estimation, sensor networks, and imaging. The effectiveness of many of these systems depend on maximizing the reception of RF power and extracting maximum information from the incident electromagnetic (EM) wave. Traditionally, this has been achieved through multiple-input multiple-output (MIMO) systems employing a spatial array of antennas that enhance the channel capacity. In this paper, we report similar increases in channel capacity obtained through the use of vector antennas consisting of co-located loops and dipoles, which can respond to more than one component of the EM field. It is shown that systems with three- and four-element vector antennas at both the transmitter and receiver operating around the frequency of 2.25 GHz support three and four times more information, respectively, as compared to conventional systems consisting of sensors with single antennas. Comparison with a simplified theoretical model of a MIMO system with co-located antennas in a rich multipath environment shows good agreement.**

***Index Terms*—Loop antenna, multiple-input multiple-output (MIMO) antenna, vector antenna.**

I. Introduction

WIRELESS systems increasingly find applications in diverse areas ranging from high-rate communication to medical imaging systems. The performance of all these systems is governed by their ability to efficiently sample the electromagnetic (EM) energy incident on them. In traditional multiple-input multiple-output (MIMO) systems, this is achieved through spatially well separated antennas that help collect the available information [1]. As an alternative, sensors consisting of multiple co-located elements responding to different components of the incident EM field (also referred to as vector antennas) can be employed, achieving the same effect as traditional MIMO systems. The increased information-theoretic capacity supported by such antennas would enable increased throughput when used for communication, increased power efficiency, better direction of arrival estimation [2], and more responsive sensors.

Manuscript received April 1, 2004; revised December 14, 2004 and February 14, 2005. This work was supported by the National Science Foundation under Grant CCF-0312696 and Grant ECS-0121389.

A. S. Konanur, S. H. Krishnamurthy, B. Hughes, and G. Lazzi are with the Department of Electrical and Computer Engineering, North Carolina State University, Raleigh, NC 27606 USA.

K. Gosalia was with the Department of Electrical and Computer Engineering, North Carolina State University, Raleigh, NC 27606 USA. He is now with General Dynamics C4 Systems (Satcom Technologies), Kilgore, TX 75662 USA.

Digital Object Identifier 10.1109/TMTT.2005.848105

Systems consisting of sensors in the form of vector antennas with co-located elements, which respond to more than one polarization, have been proposed as a means of increasing channel capacity in a rich scattering environment [3]. Research has been undertaken to theoretically account for this increase in capacity [4]–[6]. MIMO systems consisting of infinitesimal loops and dipoles [3], as well as configurations with dipoles along the sides of a cube [7] have been proposed, but there have been few instances of practical wireless systems with vector antennas built to verify the increase in information-theoretic capacity. Systems employing co-located elements consisting of three dipoles [3] and a loop and dipole [8], as well as those consisting of a loop and two dipoles in the plane of the loop [9] have been proposed.

In this paper, the properties of two such MIMO systems are investigated. The first employs a vector antenna consisting of three elements, namely, a loop and two coplanar dipoles, and operates at 2.22 GHz. The second system consists of a four-element vector antenna including a loop and three orthogonal dipoles and operates at 2.27 GHz. The properties of a MIMO communication link employing these two vector antennas are investigated experimentally and the increase in expected mutual information (EMI) is observed in a realistic scattering environment. The experimental results obtained are compared with a model [10] that is being developed to predict the effect of antenna characteristics and channel multipath environment on the mutual information and are shown to be in good agreement.

The remainder of this paper is organized in the following manner. The design and impedance characteristics of the antennas present in the system are presented in Section II. Section III provides an overview of the channel model used in simulations, and Section IV describes the experimental setup used to realize the MIMO system and obtain the channel matrices that are used in Section V to compute the mutual information and compare it with that obtained by systems employing the same number of elements in a traditional MIMO spatial array. We also compare the mutual information predicted by the model to that obtained from experimental data. Conclusions are presented in Section VI.

II. Antenna Design and Characteristics

Two MIMO systems were constructed, one with four elements and the other, coplanar, with three elements. The system with four element antennas consisted of a loop and three orthogonal dipoles. In order to separate the feeds of the four elements, the dipoles were slightly off-center fed. The three element antenna system consisted of a loop and two coplanar

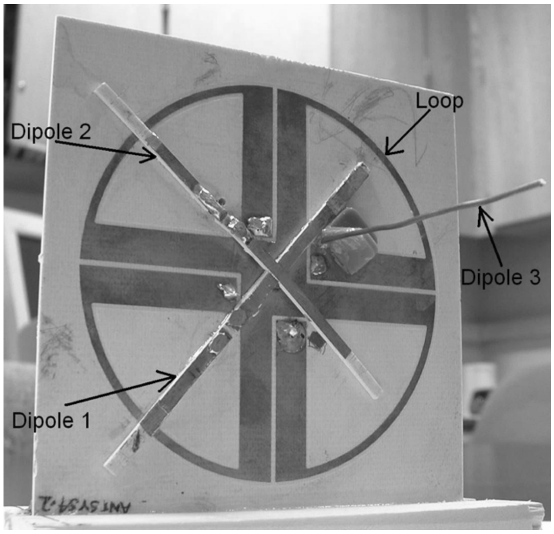


Fig. 1. Four-element vector antenna system composed of three orthogonal dipoles and a loop.

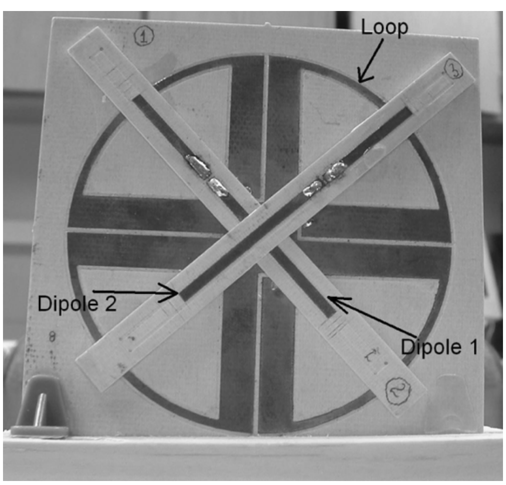


Fig. 2. Three-element vector antenna system composed of two orthogonal dipoles and a loop in a coplanar arrangement.

dipoles placed in a planar arrangement with a similar offset feeding employed for the dipoles.

Despite some preliminary research indicating that mutual coupling could help in decorrelating the multipaths [11], both antennas systems were designed with the goal of achieving low coupling amongst its elements in order to obtain high radiation efficiency.

The four- and three-element prototype antenna systems are designed to operate at the frequencies of 2.27 and 2.22 GHz, respectively. The bandwidth (i.e., S_{11} less than 10 dB) is 10 and 45 MHz, respectively. The antenna systems have both been fabricated on a 9×9 cm substrate with $\epsilon_r = 3.2$ and a thickness of 1.5 mm. Figs. 1 and 2 show the constructed antenna systems with four and three elements, respectively.

A. Loop Design

It is well understood [12] that only an electrically small loop or loop with a constant current distribution retains the radiation

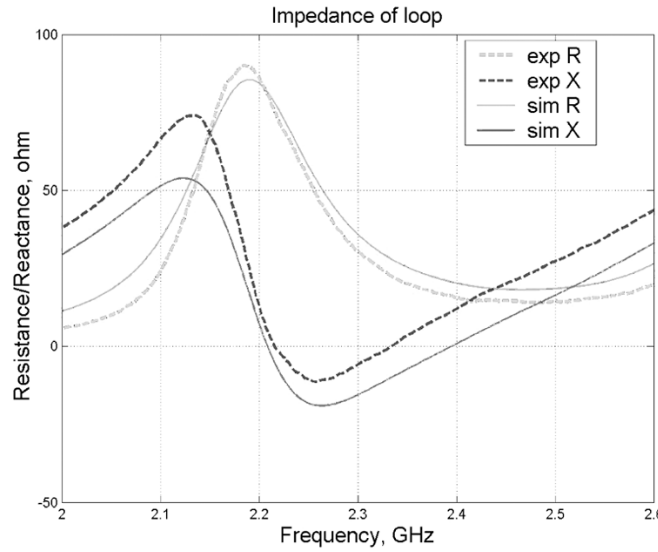


Fig. 3. Impedance of the multisector loop antenna with a ground plane: FDTD and experimental results.

characteristics of a magnetic dipole. In this study, the loop was 2-mm thick with a 7.5-cm diameter, and a relatively constant current distribution was achieved by realizing the loop by means of four pie-shaped sectors fed in phase at their corners [13]. This arrangement ensured that the currents were directed in opposite directions along adjacent feed lines, thus effectively nullifying any spurious radiation [13]. Impedance matching was achieved by feed lines of a thickness of 6 mm and a 22×22 mm square ground plane on the bottom surface of the substrate used to fabricate the loop.

The original loop design and its optimization was carried out by using an in-house finite difference time domain (FDTD) code [14] and good agreement between simulated and experimental results was found for the input impedance (Fig. 3). The radiation pattern of this multisector loop with a ground plane was simulated for comparison with that of a constant current loop, and it was verified that the E_ϕ component pattern agreed well with that of the ideal loop with a constant current. Also, the magnitude of the cross component E_θ was verified to be negligible.

B. Dipole Design

For the MIMO system with three-element antennas, the two dipoles were printed on separate substrate layers and stacked mutually orthogonally on the loop antenna. Since the loop was fed in the center, the dipoles were fed slightly off center using coaxial probes. The two dipoles were 2-mm thick and 6.5-cm long with an offset of 12 mm ($\simeq 0.08\lambda$) between the feed point and the center of the antenna system. The bottom dipole will be referred to as dipole 1, with the one on the top being dipole 2. Two notches were cut in the ground plane of the loop to isolate the dipole feeds from those of the loop and minimize coupling.

For the MIMO system with four-element antennas, the third half-wavelength dipole was made of copper wire and placed orthogonal to the other three elements through a small hole that was drilled through the substrate with an offset of 1 cm with respect to the center. It will be referred to as dipole 3 with the coplanar dipoles being dipoles 1 and 2.

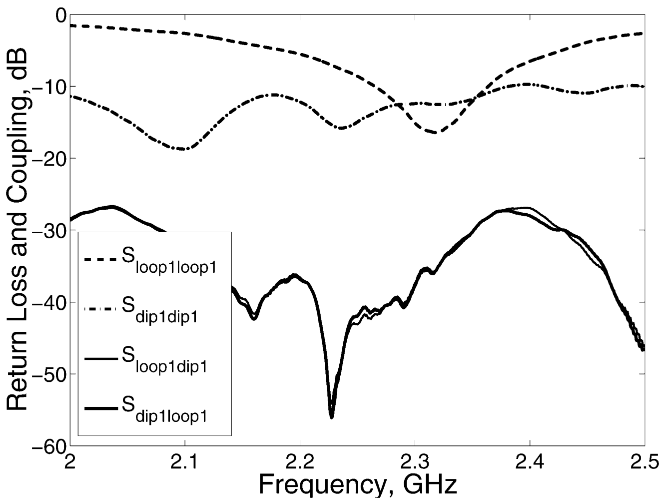


Fig. 4. Scattering parameters measured for the loop (loop1) and bottom dipole (dip1) in the four-element vector antenna case.

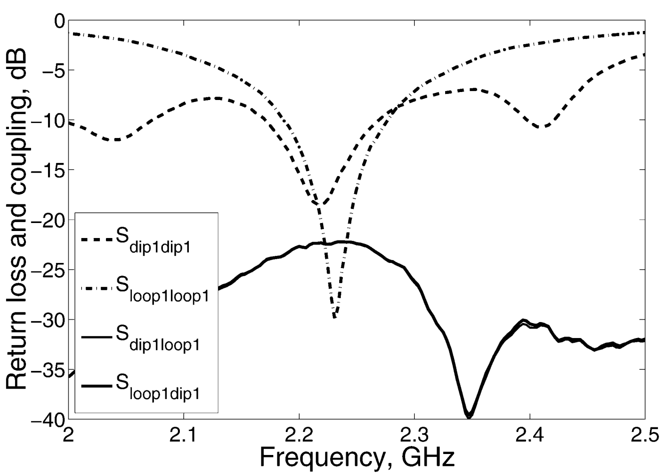


Fig. 5. Scattering parameters measured for the loop (loop1) and bottom dipole (dip1) in the three-element vector antenna case.

C. Impedance Characteristics

The placement of the loop, as well as the dipoles in close proximity necessitated the investigation of not only the matching of each antenna, but also of the mutual coupling between them. The impedance characteristics were obtained by connecting two feeds of the vector antenna to the ports of an Agilent ENA 5071B network analyzer, while terminating the remaining feeds with 50-Ω loads. As an example of the results, joint characteristics of the loop and one of the dipoles for the four-element antenna case is shown in Fig. 4. We note that both the antennas are well matched at the frequency of 2.27 GHz, with return loss being less than −10 dB and coupling lower than −26 dB over the frequency range of 2.265–2.362 GHz. Similar results are obtained for all possible pairs of these antennas with the maximum coupling being −12 dB.

The three-antenna system was designed to operate at 2.22 GHz and representative impedance and isolation characteristics of the antenna system are shown in Fig. 5. The antennas have −10-dB bandwidth between 2.186–2.269 GHz with the mutual coupling being lower than −22 dB. The maximum mutual coupling among all the antenna pairs is −19 dB. This antenna was used as a transmitting antenna with a similar antenna being used at the receiver. The complete impedance characteristics for the three-element vector antenna are presented in Appendix I.

III. Channel Model

The experimental MIMO system with vector antennas was fabricated to verify the capacity increase that was predicted by a simplified theoretical model for such antennas in a multipath environment [10]. The channel has been described in detail in [10] and here we will briefly summarize it. The model describes the joint multipath propagation of electric and magnetic fields. It captures many of the salient features of the environment of the MIMO system and yet remains tractable enough to allow for simple analysis of different vector antenna configurations, propagation conditions, and communication algorithms. The experimental observations presented in Section IV were compared with the predictions obtained from this model and, as it will be shown in Section V, found to be in good agreement.

The channel is assumed to consist of L dominant paths between the transmitter and receiver with each path resulting from scattering and reflections from physical objects placed in the far field of both the transmit and receive antennas. Thus, the waves from the source incident on the scatterer and waves from the scatterer incident on the receiver can be considered to be plane waves.

We first consider a single four-element vector antenna at the receiver. We assume that it can respond to three components of the electric field (the three dipoles respond to E_x, E_y, E_z) and one component of the magnetic field (e.g., the loop responds to H_z). We therefore consider a 4×1 column received vector $\mathbf{Y}(t) = [E_x(t), E_y(t), E_z(t), \eta H_z(t)]^T$ where the elements correspond to the complex baseband equivalents of the narrow-band signals received, η is the intrinsic impedance of the medium, and T indicates transpose operation. This received vector is the response of the antenna system to incoming plane waves, whose baseband equivalent is modeled by a 2×1 matrix $\mathbf{Z}(t)$ to account for both the horizontal and vertical components. It is assumed that the multipath component arrives at the receiver from the direction $-\mathbf{u}_r$, where

$$\mathbf{u}_r(\varphi, \theta) = \begin{bmatrix} \cos\varphi \sin\theta \\ \sin\varphi \sin\theta \\ \cos\theta \end{bmatrix} \tag{1}$$

with $\varphi \in [0, 2\pi)$ and $\theta \in [0, \pi]$ being, respectively, the azimuth and elevation of the multipath signal in receiver coordinates. For a narrow-band plane wave propagating in a nonconductive, homogeneous, and isotropic medium, the received signal can be modeled by [2]

$$\mathbf{Y}(t) = B(\varphi, \theta)\mathbf{Z}(t) + \mathbf{N}(t) \tag{2}$$

where $\mathbf{N}(t)$ represents the thermal noise and

$$B(\varphi,\theta)=\begin{bmatrix}-\sin\varphi & \cos\varphi\cos\theta\\ \cos\varphi & \sin\varphi\cos\theta\\ 0 & -\sin\theta\\ -\sin\theta & 0\end{bmatrix}. \tag{3}$$

The four-element vector antenna at the transmitter is assumed for simplicity to be a point source. The complex baseband equivalent of the transmitted signal is modeled similarly by a 4×1 matrix, $\mathbf{X}(t)=\left[E_x'(t),E_y'(t),E_z'(t),\eta H_z'(t)\right]^T$. Ignoring path losses and delays, this transmitted signal is transformed at a scatterer in the far field to $\mathbf{Z}'(t)$, where

$$\mathbf{Z}'(t)=B^T(\varphi',\theta')\mathbf{X}(t) \tag{4}$$

with φ' and θ' being the scatterer azimuth and elevation, respectively, in the transmitter coordinates and B^T being the transpose of B. At the scatterer, the plane waves undergo another transformation consisting of a change in polarization, as well as loss of amplitude. This is accounted for by a 2×2 matrix D so that $\mathbf{Z}(t)=D\mathbf{Z}'(t)$. The location and properties of the scatterer determine D. The path loss and phase shift encountered by the scattered signal before reaching the receiver may also be included in D.

In a multipath scenario consisting of L dominant scatterers, with the lth path departing the transmitter in the direction φ_l^t, θ_l^t, undergoing scattering, and arriving at the receiver from direction φ_l^r,θ_l^r, the combined signal at the receiver, assuming all the paths have approximately the same delay [10], can be described by

$$\mathbf{Y}(t)=\sqrt{\frac{\rho}{n}}H\mathbf{X}(t)+\mathbf{N}(t) \tag{5}$$

where $\mathbf{N}(t)$ is a discrete-time noise process with independent identically distributed (i.i.d.) $\mathcal{CN}(0,1)$ components, $\mathbf{X}(t)$ is constrained to unit power, ρ is the average signal-to-noise ratio (SNR) per receive antenna, $n=4$ for the four element antenna, and

$$H=\frac{1}{\alpha}\sum_{l=1}^{L}B\left(\varphi_l^r,\theta_l^r\right)D_lB^T\left(\varphi_l^t,\theta_l^t\right) \tag{6}$$

where H is the channel matrix. Here, α is the root-mean-square value of the fading path gains and ensures that the average power per receive antenna is ρ, when the inputs have unit power,

$$\alpha^2=\frac{1}{n^2}\mathcal{E}\left\{\mathrm{tr}\left[HH^\dagger\right]\right\} \tag{7}$$

where $H^\dagger$ is the conjugate transpose of H.

The modeling of the MIMO system with three-element vector antennas parallels that of the system with four-element antennas, with the main difference being in the dimensions of the received and transmitted vectors (3×1 column vectors). In this case, the received vector would be $\mathbf{Y}(t)=[E_x(t),E_y(t),\eta H_z(t)]^T$ and

$$B(\varphi,\theta)=\begin{bmatrix}-\sin\varphi & \cos\varphi\cos\theta\\ \cos\varphi & \sin\varphi\cos\theta\\ -\sin\theta & 0\end{bmatrix}. \tag{8}$$

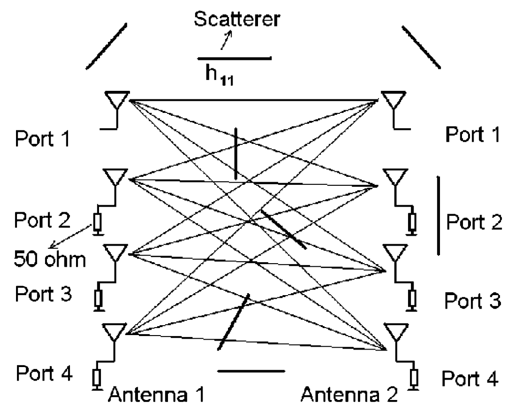


Fig. 6. Schematic representation of the experimental setup for the measurement of channel matrices H.

It is to be noted that, in the experimental setup, the scatterers were predominantly in the plane of both the transmitter and receiver. The scattering was also assumed to introduce cross polarization. Hence, D_l was assumed to have independent identically distributed $\mathcal{CN}(0,1)$ components.

IV. Experimental Setup

In order to experimentally estimate the information capacity of MIMO vector antenna systems, the fabricated antennas were employed at the transmitter and receiver in a rich scattering environment in a laboratory setting (Fig. 6). A controlled number of dominant perfect scatterers of varying sizes (up to 17 in number) were placed in the far field of the antenna system ($\mathrm{dist}\geq 2d^2/\lambda$, where d is the largest dimension in the structure i.e., 7.5 cm for the loop). The 4×4 channel matrix H was determined by measuring the transfer parameters S_{21} between each pair of elements of the transmit and receive antennas, with the remaining elements being terminated in 50-Ω loads. All the possible pairs of such ports were connected in turn to evaluate the full H matrix. The process of measuring this matrix required approximately half an hour. During this process, the environment was kept constant with personnel movement kept to a minimum. However, in order to quantify the change in the channel parameters over time, the H matrices were reevaluated so as to obtain four successive readings for the same fixed scattering environment, spanning a 2-h time frame, and it was verified that over the frequency range corresponding to the bandwidth of the antenna, the variation was less than 2%.

Forty five different channel realizations were measured. Each different channel realization included variations in the number of scatterers, their distribution in space, as well in the presence or absence of the line-of-sight. The same procedure was repeated for the three-element antenna and a set of 30 3×3 channel matrices was obtained.

V. Results

The channel matrices so obtained were all normalized by an ensemble average calculated using (7) where $n=4$ for the four-element antenna and $n=3$ for the three-element antenna. α thus ensures that the average power per receive antenna is ρ, when the inputs have unit power [10] and the SNR is ρ. The

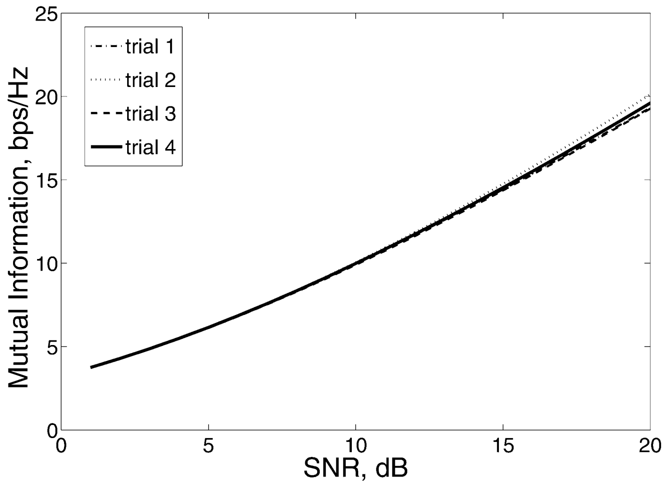


Fig. 7. Variation in mutual information over four successive sets of measurements for a fixed scattering environment with four-element antennas at both the transmitter and receiver.

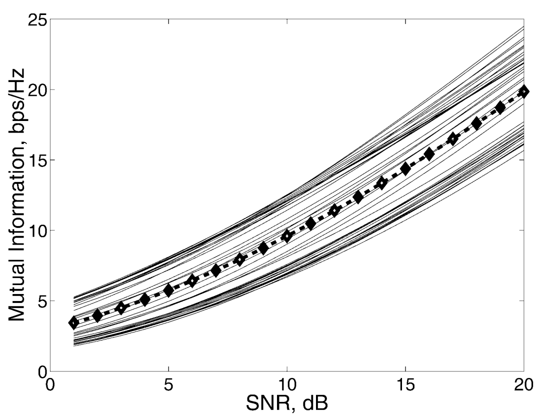


Fig. 8. Measured mutual information for MIMO system with four-element vector antennas. Line with markers is the EMI.

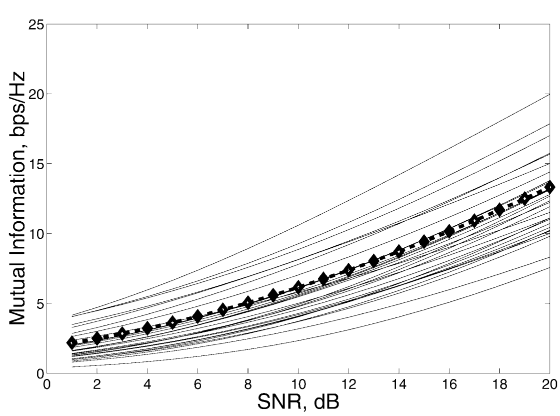


Fig. 9. Measured mutual information for MIMO system with three-element vector antennas. Line with markers is the EMI.

mutual information $I(\mathbf{Y} \wedge \mathbf{X})$ is defined for a certain SNR ρ to be

$$M(H) = \log_2 \left| I + \frac{\rho}{n} H H^{\dagger} \right| \text{ bps/Hz} \tag{9}$$

where $|\cdot|$ denotes the determinant operation.

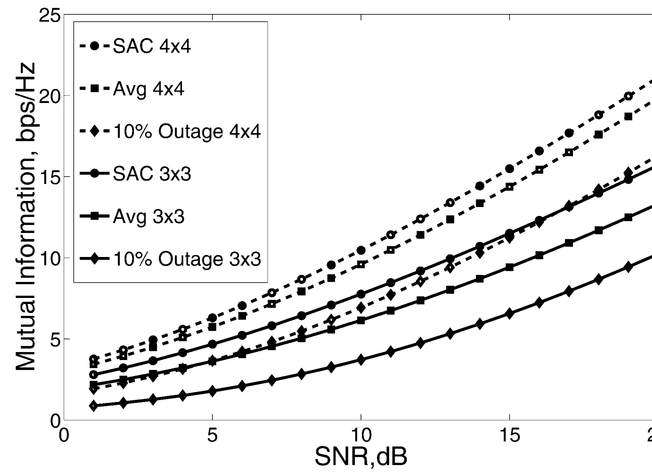


Fig. 10. EMI and 10% outage capacity of systems with four- and three-element vector antennas compared to SAC of an array with the same number of elements.

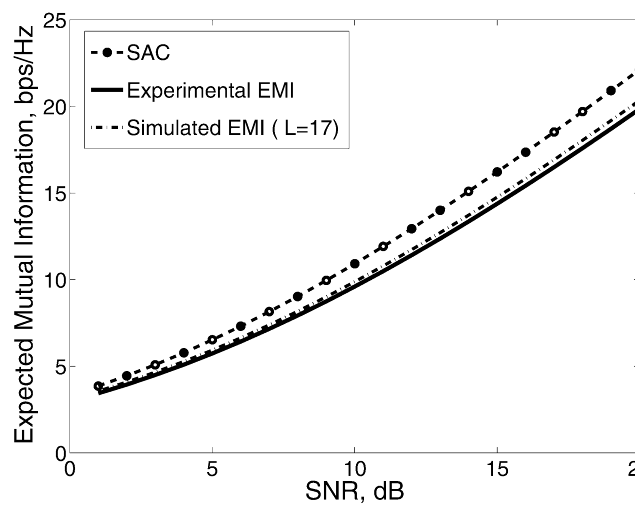


Fig. 11. Comparison between experimentally and simulated EMI for MIMO system with a four-element vector antenna with $L = 17$ scatterers.

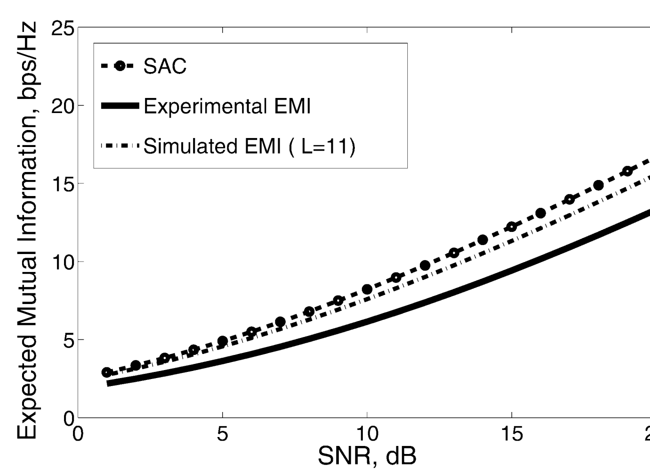


Fig. 12. Comparison between experimentally and simulated EMI for MIMO system with a three-element vector antenna with $L = 11$ scatterers.

As a further confirmation of the negligible effect of the variation of the scattering parameters' measurement over the mutual information calculation, we evaluated the mutual information using a set of four-channel matrices obtained for the same

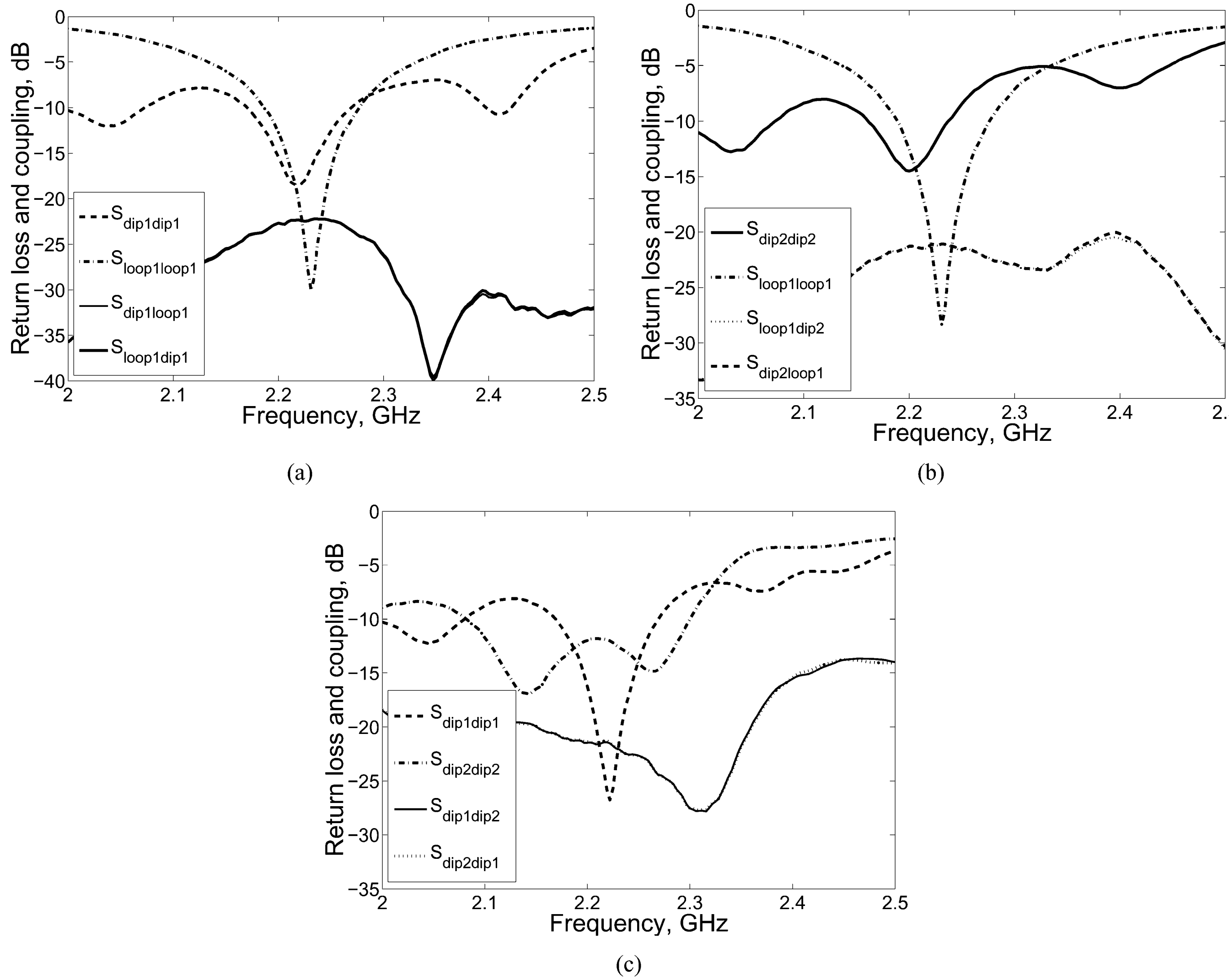


Fig. 13. Scattering parameters measured in the three-element vector antenna case. (a) Loop (loop1) and bottom dipole (dip1). (b) Loop (loop1) and top dipole (dip2). (c) Bottom dipole (dip1) and top dipole (dip2).

channel realization over a 2-h window, as mentioned in Section IV (Fig. 7). This allows us to obtain the H matrix by measuring one element at a time.

Figs. 8 and 9 show the mutual information values obtained for the systems employing the four- and three-element vector antennas, respectively. For the four-element antenna system, at 20-dB SNR, the mutual information varies from 15 to 25 bits per second/hertz (bps/Hz). In the three-element case, the mutual information at 20-dB SNR varies from 5 to 15 bps/Hz. It is interesting to note the wide variation in the mutual information, which strongly depends on the channel environment.

The EMI is obtained by averaging the mutual information $I(\mathbf{Y} \wedge \mathbf{X})$ over all channel realizations. There has been no optimization carried out with respect to the unit power input distributions and the power to each port was kept constant. For simplicity i.i.d. zero mean unit variance Gaussian variables were considered for the input distribution. Therefore, the EMI provides a lower bound on the ergodic capacity where such an optimization over the input distributions is considered. This corresponds to the case where the transmitter has no knowledge of the channel and the receiver has complete knowledge of the same. The capacity can be further enhanced by the use of the water-filling algorithm if the transmitter is fed back with the current channel state information [1]. The third quantity of interest is the 10% outage capacity, which guarantees that the throughput will remain above this level 90% of the time. It is thus the fourth lowest mutual information for the 45-channel realizations considered for the four-element vector antenna system and the third lowest mutual information for the 30 channel realizations of the three-element vector antenna system.

The EMI and the 10% outage capacity are shown in Fig. 10 for both the antenna systems under consideration. Here, the throughput in bits per second/hertz (bps/Hz) is plotted against the SNR. These are compared with the spatial array capacity (SAC) [1], which corresponds to the capacity obtained through the use of a spatially separated set of antennas at the transmitter and receiver. In the spatial array case, the elements of the channel matrix are modeled as i.i.d. complex Gaussian variables with zero mean and unit variance. The EMI curves are close to the SAC and the slopes at the high SNR are identical. Since the slope at the high SNR is proportional to the rank of H, the existence of 4 (3 for the three-element case) parallel spatial channels is demonstrated. Also, the use of the four-element antenna system leads to an improvement of close to 7 bps/Hz at 20-dB SNR over the three-element system from 13 to 20 bps/Hz.

The simulated H matrices were used to determine the EMI and compared with the experimentally determined EMI. The experimental setup for the four-element antenna case consisted of 17 principal scatterers placed predominantly on a plane perpendicular to that of the antennas ($\varphi = \pi/2$ and $3\pi/2$, $\theta = [0, \pi/3]$). We note from Fig. 11 that, for the four-element vector antenna, the simulated and experimental EMI agree closely with the values differing by 0.5 dB at 20-dB SNR. In the three-element vector antenna case, the arrangement of the scatterers was similar, but 11 scatterers were present ($\varphi = \pi/2$ and $3\pi/2$, $\theta = [0, \pi/3]$). From Fig. 12, we see that the experimentally determined EMI is lower than that predicted by the simulation by around 2 dB at 20-dB SNR. The slopes at these SNR, however, are identical confirming the scaling in capacity observed over the single-antenna case. The difference can be attributed to the simplification implicit in the model, including absence of coupling and ideal radiation patterns (of elemental loops and dipoles) considered for the elements of the antenna. This simple model, however, does provide a computationally tractable method of studying the behavior of such vector antennas in a multipath environment.

VI. Conclusion

In this paper, we have experimentally demonstrated that MIMO systems with three- and four-element vector antennas can lead to a threefold or fourfold increase, respectively, in the capacity of a communication system as compared to a conventional single-element antenna. The system with four-element vector antennas yielded an EMI of 20 bps/Hz at 20-dB SNR, while the MIMO system with three-element vector antennas is characterized by an EMI of 13 bps/Hz at the same SNR. Since these co-located vector antennas compare favorably with the SAC of a system with an identical number of transmit and receive antennas, they can be employed in lieu of ordinary MIMO systems. The favorable characteristics of these systems in being as effective as traditional spatial antenna arrays within the physical space of a single antenna could lead to their use in a number of applications, including high data-rate communication links, angle of arrival estimation, and sensors, to name a few.

Appendix
Impedance Characteristics Three-Element Antenna

Fig. 13 shows the complete return loss and coupling characteristics of the three-element vector antenna.

References

[1] E. Telatar, “Capacity of multiantenna Gaussian channels,” *Eur. Trans. Telecomm.*, vol. 10, no. 6, pp. 585–595, Jul. 1999.

[2] A. Nehorai and E. Paldi, “Vector-sensor array processing for electromagnetic source localization,” *IEEE Trans. Signal Process.*, vol. 42, no. 2, pp. 376–398, Feb. 1994.

[3] M. R. Andrews, P. P. Mitra, and R. de Carvalho, “Tripling the capacity of wireless communications using electromagnetic polarization,” *Nature*, vol. 409, pp. 316–318, Jan. 2001.

[4] A. S. Y. Poon, R. W. Brodersen, and D. N. C. Tse, “Degrees of freedom in multiple antenna channels: A signal space approach,” *IEEE Trans. Inf. Theory*, vol. 51, no. 2, pp. 523–536, Feb. 2005.

[5] T. Svantesson, M. A. Jensen, and J. W. Wallace, “Analysis of electromagnetic field polarizations in multiantenna systems,” *IEEE Trans. Wireless Commun.*, vol. 3, no. 2, pp. 641–646, Mar. 2004.

[6] S. Krishnamurthy, A. Konanur, G. Lazzi, and B. Hughes, “On the capacity of vector antenna MIMO systems,” in *Proc. IEEE Information Theory Int. Symp.*, Jul. 2004, pp. 240–240.

[7] J. Andersen and B. Getu, “The MIMO cube—A compact MIMO antenna,” in *5th Wireless Personal Multimedia Communications Int. Symp.*, vol. 1, Oct. 2002, pp. 112–114.

[8] D. D. Stancil, A. Berson, J. P. V. Hof, R. Negi, S. Sheth, and P. Patel, “Doubling wireless capacity using co-polarized, colocated electric and magnetic dipoles,” *Electron. Lett.*, vol. 38, no. 14, pp. 746–747, Jul. 2002.

[9] A. Konanur, K. Gosalia, S. Krishnamurthy, B. Hughes, and G. Lazzi, “Investigation of the performance of co-polarized, co-located electric and magnetic dipoles for increasing channel capacity,” in *IEEE AP-S Int. Conf.*, vol. 2, Jun. 2003, pp. 531–534.

[10] S. Krishnamurthy, A. Konanur, K. Gosalia, G. Lazzi, and B. Hughes, “Polarimetric multiple-input multiple-output antennas for wireless communications,” presented at the Information Sciences and Systems Conf., Baltimore, MD, Mar. 2003.

[11] T. Svantesson and A. Ranheim, “Mutual coupling effects on the capacity of multielement antenna systems,” in *IEEE Int. Acoustics, Speech, and Signal Processing Conf.*, vol. 4, 2001, pp. 2485–2488.

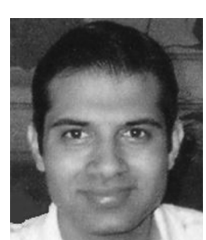
[12] C. A. Balanis, *Antenna Theory: Analysis and Design*. New York: Wiley, 1997.

[13] A. G. Kandonian, “Three new antenna types and their applications,” *Waves and Electrons*, Feb. 1946.

[14] A. Taflove, *Computational Electodynamics: The Finite-Difference Time-Domain Method*. Norwood, MA: Artech House, 1995.

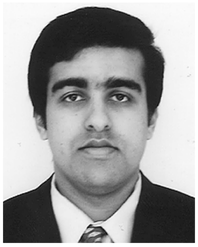
Anand S. Konanur (S’03) received the B.Tech. degree in electrical engineering from the Indian Institute of Technology, Madras, India, in 2001, and is currently working toward the Ph.D. degree in electrical and computer engineering at North Carolina State University, Raleigh.

Since January, 2002 he has been a Graduate Research Assistant with the Department of Electrical and Computer Engineering, North Carolina State University. His research interests include design of novel antennas for wireless communications, experimental channel characterization, and microwave imaging.



Keyoor Gosalia (S’01–M’05) received the B.E. degree in electronics from Sardar Patel University, Gujarat, India, in 1999, and the M.S. and Ph.D. degrees in electrical engineering from North Carolina State University, Raleigh, in 2001 and 2004, respectively.

From January 2001 to August 2004, he was a Graduate Research Assistant with the Department of Electrical and Computer Engineering, North Carolina State University. Since October 2004, he has been an RF Engineer with General Dynamics C4 Systems (Satcom Technologies), Kilgore, TX. His research interests include numerical electromagnetics, bioelectromagnetics, novel design techniques for small antennas and human body implantable antennas, planar ultra-wideband antenna systems, improving MIMO channel capacity, earth station antenna systems, microwave feed systems, and waveguide components.



Sandeep H. Krishnamurthy (S'03) was born in Shimoga, India, on November 27, 1979. He received the B.Tech. degree in electrical engineering from the Indian Institute of Technology, Madras, India, in 2001, and is currently working toward the Ph.D. degree in electrical engineering at North Carolina State University.

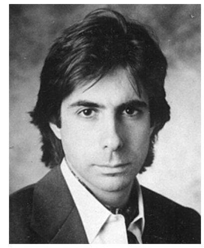
During Fall 2001, he was with the Center for Automation Research, University of Maryland at College Park. Since January 2002, he has been a Graduate Research Assistant with the Center for Advanced Computing and Communications, North Carolina State University. His research interests include information theory and signal processing for wireless communications.

Mr. Krishnamurthy is a member of Eta Kappa Nu.

Brian Hughes (S'84–M'85) was born in Baltimore, MD, on July 16, 1958. He received the B.A. degree in mathematics from the University of Maryland, Baltimore County, in 1980, and the M.A. degree in applied mathematics and Ph.D. degree in electrical engineering from the University of Maryland at College Park, in 1983 and 1985, respectively.

From 1980 to 1983, he was a Mathematician with the National Aeuronatics and Space Administration (NASA) Goddard Space Flight Center, Greenbelt, MD. From 1983 to 1985, he was a Fellow with the Information Technology Division, Naval Research Laboratory, Washington, DC. From 1985 to 1997, he was a faculty member with the Department of Electrical and Computer Engineering, The Johns Hopkins University, Baltimore, MD. In 1997, he joined the faculty of North Carolina State University, Raleigh, where he is currently Professor of electrical and computer engineering. His research interests include communication theory, information theory, and communication networks.

Dr. Hughes has served as associate editor for detection of the IEEE TRANSACTIONS ON INFORMATION THEORY, editor for theory and systems of the IEEE TRANSACTIONS ON COMMUNICATIONS, and guest editor for two special issues of the IEEE TRANSACTIONS ON SIGNAL PROCESSING. He co-chaired the 2004 Globecom Communication Theory Symposium, as well as the 1987 and 1995 Conferences on Information Sciences and Systems. He has also served on the program committees of numerous international conferences including the IEEE Global Communications Conference, the IEEE International Communications Conference, the IEEE International Symposium on Information Theory, and the IEEE Wireless Communications and Networking Conference.



Gianluca Lazzi (S'94–M'95–SM'99) was born in Rome, Italy, on April 25, 1970. He received the Dr.Eng. degree in electronics from the University of Rome "La Sapienza," Rome, Italy, in 1994, and the Ph.D. degree in electrical engineering from the University of Utah, Salt Lake City, in 1998.

He has been a consultant for several companies (1988–1994), a Visiting Researcher with the Italian National Board for New Technologies, Energy, and Environment (ENEA) (1994), a Visiting Researcher with the University of Rome "La Sapienza" (1994–1995), and a Research Associate (1995–1998) and Research Assistant Professor (1998–1999) with the University of Utah. He is currently an Associate Professor with the Department of Electrical and Computer Engineering, North Carolina State University (NCSU), Raleigh, where, from 1999 to 2003, he was an Assistant Professor. He has authored or coauthored over 80 international journal papers or conference presentations on FDTD modeling, dosimetry, and bioelectromagnetics. He is listed in *Who's Who in the World*, *Who's Who in America*, *Who's Who in Science and Engineering*, the *Dictionary of International Biographies*, and the 2000 *Outstanding Scientists of the 20th Century*.

Dr. Lazzi is an associate editor for the *IEEE Antennas and Wireless Propagation Letters*. He is the vice chair of Commission K (Electromagnetics in Biology and Medicine), U.S. National Committee of the International Union of Radio Science (URSI). He was the recipient of the 2003 ALCOA Foundation Engineering Research Achievement Award, a 2003 NCSU Outstanding Teacher Award, the 2003 NCSU Alumni Outstanding Teacher Award, a 2001 National Science Foundation (NSF) CAREER Award, a 2001 Whitaker Foundation Biomedical Engineering Grant for Young Investigators, a 1996 International Union of Radio Science (URSI) Young Scientist Award, and the 1996 Curtis Carl Johnson Memorial Award for the best student paper presented at the 18th Annual Technical Meeting of the IEEE Bioelectromagnetics Society (IEEE BEMS).