Abstract

A recent development in wireless communications is the application of multiple-input multiple-output (MIMO) systems to radio communications via use of multiple antennas. In order to investigate the technology’s potential, an experimental MIMO system which contains four-element antenna arrays (4 × 4) has been developed at the University of Alberta. The system is used to obtain MIMO channel measurements in a typical indoor office environment in the ISM band (902 - 928 MHz). Measurement campaigns were performed using different antenna spacings and two different types of antenna: half-wavelength (λ/2) centre-fed dipole and dual polarized patch. The measurements are used to calculate the theoretical channel capacities for a 4 × 4 MIMO system. The measurements confirm the high capacity potential of a MIMO channel, with ergodic capacity of approximately 21 bits/channel use available with either antenna type, when the antenna element separation is λ/2 or larger at a signal-to-noise ratio of 20dB. An introduction to basic MIMO theory, a discussion of the University of Alberta wireless MIMO testbed, and observations regarding the measured indoor MIMO channel are presented in the paper.

1 Introduction

MIMO (multiple-input multiple-output) wireless systems using arrays of antennas instead of single transmit and receive antennas, hold the promise of providing data rates far exceeding those of conventional wireless systems [1]. Such MIMO systems operate by transmitting multiple signals in the same frequency band and at the same time over multiple transmit antennas. At the receiver, multiple antennas are used and the received signals are processed to separate the different transmitted data streams.
MIMO communications is made possible by the extension of receiver processing to include spatial dimensionality as well as time (spatio-temporal processing). Multiple propagation paths, as occur in a scattering propagation environment, become distinguishable at the MIMO receiver due to small differences in arrival times and can be used to carry additional information across the channel [2]. Although the idea of using multiple antennas is not new – it finds application in such techniques as diversity combining and beam-steering - only MIMO systems approach the information carrying capacity potential of such channels.

It is likely that future applications of MIMO technology include both fixed and mobile uses for indoor as well as outdoor environments. An example of a potential mobile application for MIMO communications has recently been studied in [3], which describes an investigation on using the technology for high speed (around 1 Mb/s) communication with subway trains. A MIMO system for residential fixed wireless access was developed by Iospan Wireless (now part of L-3 Communications Inc.) [4]. Another company, Antenova Ltd, is examining the use of MIMO for laptops and handheld consumer devices [5]. A MIMO chipset has already been developed by Airgo Networks, a start-up company in Silicon Valley [6].

A MIMO system with $N_t$ transmit antennas and $N_r$ receive antennas utilizes a channel with $N_t N_r$ separate paths and the gains for these paths are described using an $N_t \times N_r$ matrix $H$. The information theoretic capacity, i.e., the maximum throughput of a MIMO channel, is dependent upon the characteristics of its transmission matrix $H$, in particular its singular values. As shown in Section 2, these depend on the propagation environment, which ultimately determines the capacity of MIMO channels. In general, a rich scattering environment as occurs in indoor wireless communications, tends to produce channels with high capacity.

Since wireless channels are inherently complex and difficult to characterize, obtaining channel measurements is the only accurate and reliable method of determining the yet not fully understood capacity potential of such MIMO systems. In order to measure the capacity of MIMO systems in typical propagation environments, a flexible and mobile MIMO testbed for up to $4 \times 4$ arrays has been developed in the iCORE HCDC (High Capacity Digital Communications) laboratory at the University of Alberta, Canada.

In this paper, we present results of several MIMO channel measurement campaigns for non-line of sight channels in an indoor office environment, using two different types of antennas: standard $\lambda/2$ dipole and specially designed dual-polarized patch antennas [7]. The singular values of the measured channel gain
matrices are calculated and analyzed along with the channel capacities.

In Section 2 of this paper we discuss the capacity formulas for MIMO channels and explain the impact of system and propagation environmental parameters. We then discuss implementation issues along with the need to perform MIMO channel measurements in Section 3. Section 4 gives a description of our testbed and our system validation tests. In Section 5, we show how our campaigns were conducted and present the results for them. Finally, Section 6 gives conclusions and outlines planned future research.

2 The Multi-Antenna MIMO Channel

2.1 Channel Model and Capacity

A MIMO transmission system uses $N_t$ transmit and $N_r$ receive antennas. Each antenna transmits symbols from a complex symbol alphabet each with power $E_s$, per signaling interval, such that $\sum_i E_{si} = E_s$ is constant for each use of the channel. These transmit symbols are modulated by a suitable pulse waveform, up-converted to the desired transmission band, and sent over the $N_t$ transmit antennas. The signals from the receive antennas are mixed down to baseband, sampled, and fed into the decoder. A basic block-level diagram of a MIMO system is shown in Figure 1.

The wireless transmission channel is a linear channel, and, assuming that timing recovery has been accomplished, the received sampled signal $y_{jl}$ for the $j^{th}$ receive antenna at time $l$ is given by

$$y_{jl} = \sum_{i=1}^{N_t} \sqrt{E_{si}} h_{ij} c_{il} + \eta_{jl} \quad (1)$$

where $\eta_{jl}$ is a sample of circularly symmetrical complex Gaussian noise with variance $N_0$, $c_{il}$ is the sampled transmitted signal and $h_{ij}$ is the complex path strength from transmit antenna $i$ to receive antenna $j$. It contains all linear effects on the signal, such as propagation power loss and phase shifts, fading due to multipath, cross-talk, antenna coupling, and polarization. We have furthermore assumed that the symbol rate is low enough such that frequency selectivity caused by time-of-arrival differences between the various multipath replicas of the received signal is not a issue that manifests itself noticeably. This implies symbol rates of about 10 Mbaud or less for indoor transmission, and about 1 Mbaud or less for outdoor situations.
The entire MIMO channel can now succinctly be characterized by the linear algebraic relationship

\[ y = HA+a + n, \quad A = \begin{bmatrix} \sqrt{E_{s1}} \\ \sqrt{E_{s2}} \\ \cdot \cdot \cdot \\ \sqrt{E_{sN_t}} \end{bmatrix} \]  

(2)

The information theoretic capacity of the discrete channel in (2) can be calculated from basic information theoretic concepts \[8\] as

\[ C_I = \log_2 \det \left( I + \frac{\rho}{N_t} HEH^+ \right) \text{[bits/channel use]}, \]

(3)

where \( \rho = \frac{E_s}{N_0} \) is the total symbol signal-to-noise ratio,

\[ E = \frac{1}{E_s} \begin{bmatrix} E_{s1} \\ E_{s2} \\ \cdot \cdot \cdot \\ E_{sN_t} \end{bmatrix} \]

(4)

and \( H^+ \) is the conjugate transpose of \( H \). Since the channel parameters are time-varying, \( C_I \) is interpreted as the “instantaneous” channel capacity for a given channel realization \( H \). For a time-varying channel this capacity has to be averaged to calculate the ergodic channel capacity \( C = E_H(C_I) \). Telatar \[9\] has presented closed form solutions for \( C \) in the case where the \( h_{ij} \) are independent complex Gaussian fading channel gains.

The channel matrix \( H \) can be decomposed via the singular value decomposition (SVD) method \[10\] into the product

\[ H = UDV^+ \]

(5)

where \( U \) and \( V \) are unitary matrices, i.e., \( UU^+ = I \), and \( VV^+ = I \). The matrix \( D \) contains the singular values \( \{d_n\} \) of \( H \) on its diagonal, which are the positive square roots of the non-zero eigenvalues of \( HH^+ \) or \( H^+H \). Note that \( D \) may not be a square matrix, which simply means that the number of non-zero singular values can be no larger than the minimum size of the matrix, and is in fact equal to its rank. The SVD allows us to write the channel equation in an equivalent form

\[ y = UDV^+Ac + n \]

\[ U^+y = \tilde{y} = D\tilde{c} + \tilde{n} \]

(6)

where \( \tilde{c} = V^+Ac \). This leads to parallel Gaussian channels \( \tilde{y}_n = d_n\tilde{x}_n + \tilde{n}_n \) as shown in Figure 2.
The capacity of the MIMO channel is now determined by the well-known waterfilling theorem \[11\] as

\[
C_I = \sum_{n=1}^{N} \log \left( 1 + \frac{d_n^2 E_n}{N_0} \right) = \sum_{n=1}^{N} \log \left( \frac{d_n^2 \mu}{N_0} \right)
\]

(7)

where \(E_n\) is the power allocated to channel \(n\), and \(N = \min(N_r, N_t)\) is the rank of \(H\) excluding pathological cases. This capacity is achieved with the waterfilling power allocation:

\[
\frac{N_0}{d_n^2} + E_n = \mu; \quad N_0/d_n^2 < \mu
\]

(8)

\[
E_n = 0; \quad N_0/d_n^2 \geq \mu
\]

(9)

If the channel is known at the transmitter, the signal strategy shown in Figure 2, which is based on the SVD, achieves capacity by transforming the channel into a set of parallel channels.

However, channel knowledge is not typically available at the transmitter, and the only choice we have is to distribute the energy uniformly over all component channels. This leads to the Symmetric Capacity

\[
C_{sym} = \sum_{n=1}^{N} \log \left( 1 + \frac{d_n^2 E_s}{N_t N_0} \right) = \log \prod_{n=1}^{N} \left( 1 + \frac{d_n^2 E_s}{N_t N_0} \right)
\]

(10)

Using the fact \(\det(M) = \prod \lambda(M)\), and \(\det(I + M) = \prod (1 + \lambda(M))\) we obtain the equivalent formulations:

\[
C_{sym} = \log \prod_{n=1}^{N} \left( 1 + \frac{d_n^2 E_s}{N_t N_0} \right) = \log \det \left( I_{N_r} + \frac{\rho}{N_t} HH^+ \right) = \log \det \left( I_{N_r} + \frac{\rho}{N_t} H^+ H \right)
\]

(11)

Fundamentally, the capacity of a MIMO channel is governed by the singular values of \(H\) which determine the channel gains of the independent equivalent parallel channels.

2.2 Capacity Potential and Limitations of MIMO Channels

If the channel paths \(h_{ij}\) are uncorrelated, most channel realizations are of high rank with the eigenvalues of \(HH^+\) distributed according to a Wishart distribution \[9\]. The case of equal singular values represents a limiting case, for which the capacity behavior is given by

\[
C_{\text{high}} = \sum_{n=1}^{N} \log \left( 1 + \frac{d_n^2 \rho}{N_t} \right) \approx N \log \left( 1 + \frac{N_r \rho}{N_t} \right)
\]

(12)

and capacity increases linearly with the minimum of the number of elements in either of the two antenna arrays. On the other hand, if component channels are completely correlated, such as occurs in scatterfree long-distance wireless connections, (e.g. in a satellite-ground radio link) all rows \(h_j\) of \(H\), the array
response vectors, are approximately equal, and $H$ is a scaled version of the all-ones matrix of dimension $N_r \times N_t$, whose only non-zero singular value is $d_s^2 = N_t N_r$. As a result

$$C_{\text{low}} \approx \log (1 + N_t \rho)$$  \hspace{1cm} (13)$$

In this case the channel capacity grows only logarithmically with the number of (receive) antennas. Real-world situation will lie somewhere between these two extremes, with the capacity determined by the complex propagation environment in which the system has to function. This leads to the necessity to carefully analyze and measure such candidate environments to obtain precise values. In fact, for scatter-free propagation the array response vectors become correlated very rapidly (see also [12]) and the channel matrix $H$ looses rank, turning the MIMO channel into nothing more than a fancy conventional wireless link.

Furthermore, the capacity potential of a MIMO channel requires relatively high-energy transmitters as can be shown as follows: if the signal-to-noise ratio $\rho$ is low, a Taylor Series approximation of $\log(1+x) \approx x$ for small $x$ lets us develop both (12) and (13) as

$$C_{\text{high}} \approx N \frac{N_r \rho}{N_t}; \hspace{1cm} C_{\text{low}} \approx N_r \rho$$  \hspace{1cm} (14)$$

which now both grow linearly with the array sizes. This indicates that correlation in the channel has no effect on capacity for low SNR values and $N_t \leq N_r$. The sole effect of an increased number of antennas is that of gathering more received power. Communications is fundamentally power limited and the additional dimensionality offered by a “high-rank” MIMO channel cannot be exploited.

From the physical model it is evident that the geometry of the propagation environment plays a significant role. The transmission strategy for both low-rank MIMO channels as well as low SNR MIMO channels is identical and straightforward. Concentrate all transmit power on a single antenna and use maximum-ratio combining of the $N_r$ receive antennas to feed a single receiver. Of course, the transmit antenna array can be used to shape the transmission beam to direct power to the desired receiver. These are the well-known traditional beam steering techniques.

## 3 MIMO Channels in the Real World

In order to better understand MIMO channels in real-world environments, accurate MIMO channel measurements are needed. One issue with wireless channels is the large power variations that occur in the
received signals due to multipath fading if one or both of the terminals are mobile. Figure 3 is a plot of the received signal obtained from measurements with the HCDC MIMO testbed for a single transmit-receive antenna pair in an indoor office environment without automatic gain control. It illustrates the rich multipath environment, which results in power fluctuations. Clearly visible are the signal fades approximately every wavelength, i.e., every 30cm. If a receiver moves through this signal field, the spatial fading pattern turns into time-varying signal fading. As can be appreciated, channel tracking under such conditions is a major challenge for MIMO systems, where there are $N_r \times N_t$ such channels which vary largely independently. Channel tracking, however, is essential for MIMO communications since the complex channel gain information is needed to separate the co-existing signals. In the absence of channel tracking, orthogonal space-time codes [13], or differential space-time modulation [14] has to be used, both severely limiting the achievable maximal data rates.

Fortunately, and at first glance perhaps paradoxically, despite the power fluctuations that occur in the received signals, the MIMO channel capacity itself is very stable. This is due to the fact that the large number of component channel gains tend to average out, presenting an “average” MIMO channel to the receiver at all times. In other words, it is very unlikely that all of the channel paths will be in a deep fade at the same time. The channel capacity of the $4 \times 4$ MIMO channel, one of whose components is the fluctuating signal in Figure 3, is shown in Figure 4. Astonishingly, the channel capacity fluctuates with a standard deviation of less than 10% of the average.

Apart from the fundamental problem of acquiring precise timing for the symbols, which we shall not discuss in this paper, MIMO receivers face the task of demodulating potentially very large signal sets. As an example, consider a $16 \times 16$ MIMO system with 8PSK modulation employed on each transmit antenna. Each space-time symbol will therefore consist of $3 \times 16 = 48$ bits with $2^{48}$ making approximately $10^{15}$ signal points. In order to handle such large symbol sets, maximum likelihood demodulation becomes infeasible and advanced demodulation techniques, such as signal layering [15], are required.

4 Testbed Description

The mobile MIMO measurement system developed at the University of Alberta consists of independent transmitter and receiver stations. Each station is comprised of an FPGA (field programmable gate array) development board for baseband processing, a custom RF (radio frequency) module (for up-, or down-
conversion) and a custom antenna array structure. A PC is used to process captured data from the receiver FPGA board. A block diagram of the hardware components is shown in Figure 5.

4.1 The Signal Processing Side

All signal processing functions are realized on GVA-290 FPGA development boards manufactured by GV and Associates Inc. Each GVA-290 board contains four Analog Devices AD9762 DACs, four Analog Devices AD9432 ADCs, two Xilinx Virtex-E FPGAs and a Xilinx Spartan-II FPGA.

Four-channel RF modules were custom built for this project by SignalCraft Technologies Inc. Each transmit module receives the spread signals at an IF of 12.5 MHz and up-converts them to the ISM band (902-928 MHz). The signals can be amplified to a maximum output power of 20 dBm per channel. The receive module mirrors the functions of the transmit module. The four signals received from the antennas are amplified and down-converted from the ISM band to 12.5 MHz. On both sets of RF modules, only one local oscillator is used at each frequency stage to generate the signals that mix with the four input signals. This ensures that no phase shifts are introduced between the four channels by the modules, and that the different received signals do not rotate with respect to each other.

An FPGA image has been developed for the transmitter station which generates and simultaneously transmits four filtered 500 kchips/s spread spectrum signals (see Figure 6). The spreading sequences used are orthogonal Walsh codes of length 32 multiplied by a pseudorandom $m$-sequence [16] to provide nearly ideal correlation properties. The chip pulse has a truncated square-root raised-cosine shape with a cutoff frequency of 500 kHz and a roll-off factor of 0.31. The four filtered signals are up-converted digitally to an IF of 12.5 MHz before being sent to the DACs.

At the receiver station (see Figure 7), all of the signal processing is performed on the FPGA development board. The four parallel down-converted signals are received at an IF of 12.5 MHz from the RF receive module. The signals are sampled simultaneously by the ADCs at 50 Msps and passed into an FPGA. In the FPGA, the signals are digitally down-converted to baseband then decimated to 1 Msps (2 samples/chip). Subsequently, they are convolved through a square-root raised-cosine matched filter with the same characteristics as the modulation pulse shaping filter. An involved parallel synchronization algorithm [17] is used to synchronize the reference signals in the receiver with the received signals. Each received signal is correlated with stored copies of the four PN-spread signals that were used in the transmitter. The
outputs of the correlation operation create complex $4 \times 4$ correlation matrices for each sample instant. The individual matrix elements are squared and subsequently all 16 values are added together.

Despite the very good autocorrelation properties of the spreading sequences used, due to the random phases of the component channels, the composite autocorrelation function at the receiver can have significant side lobes. A plot of the summed squared correlation values for equal gains is shown in Figure 8 for one symbol period. (the period of the spread signals is 64 $\mu$s). Correct identification of the main correlation peak establishes the symbol timing reference. The correlation matrix at the selected synchronization instant is equal to the MIMO channel gain matrix $H$. These channel gain matrices are sent from the FPGA board to a Matlab program on a PC where they are used to calculate the singular values and calculate channel capacity.

4.2 The Electro-Magnetics Side

Two types of antenna arrays are used with our testbed. One array type is comprised of four ($\lambda/2$) centre-fed dipoles and the other consists of two dual-polarized patch antennas [7]. The centre-fed dipoles are created by mounting $\lambda/4$-length monopole antennas onto conductive sheets. For both array cases, the antennas were placed in a row so as to make a broadside array. Each patch antenna contains two co-located elements with orthogonal polarizations.

Both antenna types have potential advantages and disadvantages with respect to increasing the capacity of the MIMO channel. Performing identical measurement campaigns with both array types allows the determination of which antenna type yields higher channel capacity.

The patch antenna array has the potential advantage of employing two orthogonal polarizations while the dipole antenna array uses only a single polarization. This means that in a $4 \times 4$ MIMO system with patch antennas, ideally each transmitted signal is affected by only one significant interfering signal. It is then natural to conclude that reduced interference would lead to a higher MIMO channel capacity. However, the scattering that will occur in a multipath radio environment results in rapid loss of polarization separation.

The dipole antenna has the potential advantage of having a uniform 360 degree radiation pattern in the plane perpendicular to itself, whereas the patch antenna has a radiation pattern that is nearly uniform over less than 180 degrees in its perpendicular planes. The dipole antenna’s wider radiation pattern leads one to expect it to provide a higher capacity since it creates a richer multipath environment. The transmitted
signals from the dipole antenna array radiate in all directions resulting in reflections and scattering off more objects. Also, the receiver antennas would detect incoming waves from all directions instead of only from a hemisphere.

A consideration concerning the link budget is the efficiency of the antenna: what fraction of the electrical energy that enters the antenna connector actually radiates from the antenna as electromagnetic energy. Energy is lost due to antenna dielectrics and protective covers. A rough measurement of the efficiency of an antenna can be obtained using an unobstructed line-of-sight channel. For such a channel, the theoretical received power can be calculated using the transmission formula [18]. A comparison of the measured received power to the theoretical received power (assuming perfectly efficient antennas) will determine the antenna efficiencies.

Efficiency measurements have been performed for both the dipole and patch antennas. Our testbed has been set up with one transmitter antenna and one receiver antenna. The transmitter and receiver stations have been placed a few metres apart. An efficiency of 74% for the dipole antenna and an 18% efficiency for the patch antenna have been measured and calculated. Thus, the dipole antenna has a clear advantage over the patch antenna if the link power budget is an issue.

In order to verify that the polarization separation of our patch antenna arrays is good, experiments have been conducted to measure the cross polarization discrimination. The experiment measures the amount of power emitted in one polarity that is received by the antennas in the cross polarity. The experiment is performed by placing the two antennas of the receiver station very close (60 to 65 cm separation) to the antennas of the transmitter array. This separation is large enough to avoid near field electromagnetic effects but small enough to create a strong line-of-sight path with virtually no multipath effects.

Using this testbed set-up, each of the four transmitter signals is radiated separately from its corresponding antenna while the power at all four received antennas is measured. Table 1 shows the measured relative power of the received signals with cross polarization. As can be observed in the table, the powers of the received cross-polarized signals are 11.9 dB to 17.1 dB below the power of their corresponding received co-polarized signal. Although our measured cross polarization discrimination values are not as large as when the antennas are used in an ideal transmission environment [7], they are still large enough to ensure that cross-coupling will have a minimal effect on our channel gain measurements.
In order to assess the error in the channel gain measurements obtained with our system, measurements have been made for an uncoupled line-of-sight (LOS) MIMO channel [19]. This LOS channel was created by removing the antenna arrays and connecting the four RF outputs from the transmitter to the four RF inputs of the receiver with cables. For this uncoupled LOS channel, the diagonal elements \( h_{ii}, i = 1, \ldots, 4 \) of the channel gain matrix correspond to the connected paths and should have the same magnitude. All the off-diagonal values \( h_{ij}, i \neq j \) represent the non-existent cross paths and should be zero. The deviation in the measured values from this expected result represents the system error.

The averaged normalized results from 80 measurements are shown in Table 2 and demonstrate that the error introduced by the system is low. Measured gains of all connected paths are less than 0.5 dB of each other. The power measured in each of the non-existent cross channels is at least 27 dB below the power measured for connected paths.

5 Measured Capacity Results

5.1 Details of Measurement Campaigns

The fifth floor of the ECERF (Electrical and Computer Engineering Research Facility) building on the University of Alberta campus was selected as the location for the indoor channel measurement campaigns. The ECERF building is constructed of steel, concrete and drywall which is typical of many modern office buildings.

Four campaigns were performed in total. Three of the campaigns used antenna arrays of dipole antennas and one used dual polarized patch antennas. In all cases, the antennas at each station were placed so as to create a broadside array. For the three dipole antenna campaigns, the four antennas were spaced at distances of \( \lambda/8 \), \( \lambda/4 \) and \( \lambda/2 \) at both stations. For the patch antenna campaign, the two microstrip patch elements were spaced \( \lambda/2 \) apart.

Once we had obtained channel gain data in our measurement campaigns, the matrices were normalized according to

\[
H = G \sqrt{\frac{N_r N_t}{||G||^2}}
\]

where \( ||G||^2 \) is the squared Frobenius norm of \( G \), i.e., the sum of the squared magnitudes of the elements of \( G \). The purpose of normalizing the gain matrices is to eliminate the effect of propagation power loss.
From $H$ the MIMO channel capacities were calculated using equation (3) with $\rho$ set to 20 dB.

For comparison, $4 \times 4$ MIMO channel matrices have also been generated by simulation. Each element of each simulated matrix is created using an independent complex Gaussian random number generator with zero mean and unit variance per dimension. The capacities of the simulated channels have been calculated using (3) after being normalized. Channel gain matrices generated in this fashion have been used in several MIMO capacity studies [20].

5.2 Channel Measurements and Comparisons

Figure 9 shows the complementary cumulative distribution function (CCDF) curves for the three dipole antenna campaigns along with the simulated Gaussian case [21]. Two observations can be made by comparing the curves. First, the mean capacity of the simulated Gaussian channel is about 2 bits/channel use higher than the highest mean capacity of the dipole configuration. This difference may be due to a small specular component present in the MIMO channel. Nevertheless, the measurements clearly demonstrate the high capacity potential. The second observation is that the capacity drops as the antennas are placed closer together, which is not surprising since greater correlation will occur between the received signals.

Figure 10 shows the CCDF curves for the $\lambda/2$ spaced dipole and $\lambda/2$ spaced patch antenna cases [22]. A slightly higher capacity can be observed for the patch antenna array case. The CCDF curve for the $\lambda/2$ spaced dipole array in Figure 10 is slightly different from the curve for the $\lambda/2$ spaced dipole array in Figure 9. This is because different locations were used for the transmitter antenna array.

Another meaningful comparison is that of the singular values (SV) of the channel gain matrices for the four different measurement campaigns. MIMO channels which exhibit higher channel capacity have singular values which are approximately equal while channels with lower capacity have fewer dominant singular values. Figure 11 shows the histogram of the measured SVs for the dipole antenna array campaign when $\lambda/2$ element spacing is used. The solid curve shows the theoretical Wishart distribution for an independent Gaussian channel with twice as many receive as transmit antennas (i.e. with a number of significant SVs which is half the size of the array). This curve provides a good match and would suggest that there is about a rank two contribution from diffuse transmission components, while the second mass of SVs with large values around 3-3.5 result from specular transmission paths.
6 Conclusion and Future Outlook

As shown in Section 5, the mean MIMO channel capacities that we have obtained for the two \( \lambda/2 \) separation cases are 21.3 bits/channel use (patch antenna case) and 21.1 bits/channel use (dipole case). It is worthwhile to compare these values to those derived from the limiting case equations (12) and (13). For \( \rho = 20 \) dB and \( N_t = N_r = 4 \), we can calculate \( C_{\text{high}} = 26.6 \) bits/channel use and \( C_{\text{low}} = 8.6 \) bits/channel use. Thus, the mean capacities for our typical indoor office environment reach approximately 80 percent of the ideal upper limit case.

A better appreciation of the potential data rate increase can be obtained with a different explanation of the upper limiting case equation (12). The capacity potential that is defined by this equation equals the capacity of \( N_t \) separate orthogonal channels, e.g., different frequency bands. Thus, with a \( 4 \times 4 \) system using a 1 MHz frequency band, we obtain 80\% of the capacity of single-antenna links using four 1 MHz bands.

A second interesting conclusion that can be drawn from our results is that the CCDF curve for the \( \lambda/2 \) spaced patch antenna case is slightly higher than the CCDF curve for the \( \lambda/2 \) spaced dipole antenna. Therefore, for our measurement location, the advantage of using dual polarized signals completely compensates the disadvantage of having antennas with a partial (hemispheric) radiation pattern.

There are several intriguing extensions to the work that we have presented here. One is to obtain channel measurements for a much larger system (e.g. \( N_t = N_r = 16 \)) to determine if the same throughput potential relative to the upper limit case exists, i.e., if there is enough scattering in the channel for large capacity gains. This question is undoubtedly dependent on the particular transmission environment (indoor, outdoor, indoor-outdoor, etc.), and measurements are currently under way. Finally, it will be valuable to expand our system to transmit data and establish how much of the MIMO channel capacity is achievable in a practical system. We are currently pursuing these research avenues.
References


[17] patent pending.


7 Figures and Tables

<table>
<thead>
<tr>
<th>Radiating Transmitter Antenna</th>
<th>Relative Power Cross Path 1 (dB)</th>
<th>Relative Power Cross Path 2 (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>-12.9</td>
<td>-12.3</td>
</tr>
<tr>
<td>2</td>
<td>-17.1</td>
<td>-11.9</td>
</tr>
<tr>
<td>3</td>
<td>-17.1</td>
<td>-14.7</td>
</tr>
<tr>
<td>4</td>
<td>-16.9</td>
<td>-12.0</td>
</tr>
</tbody>
</table>

Table 1: Measured cross polarized power in the patch antenna arrays (relative to the main co-polarized path)

<table>
<thead>
<tr>
<th>Channel Path</th>
<th>Power (dB)</th>
<th>Channel Path</th>
<th>Power (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$h_{11}$</td>
<td>-0.33</td>
<td>$h_{31}$</td>
<td>-30.13</td>
</tr>
<tr>
<td>$h_{12}$</td>
<td>-34.23</td>
<td>$h_{32}$</td>
<td>-32.32</td>
</tr>
<tr>
<td>$h_{13}$</td>
<td>-37.73</td>
<td>$h_{33}$</td>
<td>0.00</td>
</tr>
<tr>
<td>$h_{14}$</td>
<td>-32.62</td>
<td>$h_{34}$</td>
<td>-34.33</td>
</tr>
<tr>
<td>$h_{21}$</td>
<td>-29.40</td>
<td>$h_{41}$</td>
<td>-30.71</td>
</tr>
<tr>
<td>$h_{22}$</td>
<td>-0.16</td>
<td>$h_{42}$</td>
<td>-31.84</td>
</tr>
<tr>
<td>$h_{23}$</td>
<td>-31.22</td>
<td>$h_{43}$</td>
<td>-27.63</td>
</tr>
<tr>
<td>$h_{24}$</td>
<td>-37.47</td>
<td>$h_{44}$</td>
<td>-0.26</td>
</tr>
</tbody>
</table>

Table 2: Average squared path gain values for an uncoupled LOS channel

![Figure 1: Basic communication system for MIMO channels.](image-url)
Figure 2: Optimal signal processing if the channel is known at the transmitter.

Figure 3: Sample received power profile for an indoor office environment
Figure 4: Sample $4 \times 4$ MIMO channel capacity profile for an indoor office environment: $\rho = 20$ dB
Figure 5: Block diagram of the MIMO testbed hardware

Figure 6: Block diagram of the transmitter signal processing chain (single channel)
Figure 7: Block diagram of the receiver signal processing chain
Figure 8: Summed squared correlation response for one code sequence period
Figure 9: CCDF of the measured channel capacity: $\rho = 20$ dB
Figure 10: CCDF of the measured channel capacity: $\rho = 20$ dB
Figure 11: Histogram of SVs from the $\lambda/2$ separated dipole measurements