The Impact of Antenna Diversity on the Capacity of Wireless Communication Systems

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***Abstract*— For a broad class of interference-dominated wireless systems including mobile, personal communications, and wireless PBX/LAN networks, we show that a significant increase in system capacity can be achieved by the use of spatial diversity (multiple antennas), and optimum combining. This is explained by the following observation: for independent flat-Rayleigh fading wireless systems with N mutually interfering users, we demonstrate that with $K+N$ antennas, $N-1$ interferers can be nulled out and $K+1$ path diversity improvement can be achieved by each of the N users. Monte Carlo evaluations show that these results also hold with frequency-selective fading when optimum equalization is used at the receiver. Thus an N-fold increase in user capacity can be achieved, allowing for modular growth and improved performance by increasing the number of antennas. The interferers can also be users in other cells, users in other radio systems, or even other types of radiating devices, and thus interference cancellation also allows radio systems to operate in high interference environments. As an example of the potential system gain, we show that with 2 or 3 antennas the capacity of the mobile radio system IS-54 can be doubled, and with 5 antennas a 7-fold capacity increase (frequency reuse in every cell) can be achieved.**

I. INTRODUCTION

The chief aim of this paper is to demonstrate theoretically that antenna diversity (with optimum combining) can substantially increase the capacity of most interference-limited wireless communication systems. We also study implementation techniques and issues for achieving these increases in operating systems. Increasing the number of users in a given bandwidth is the dominant goal of much of today's intense research in mobile radio, personal communication, and wireless PBX/LAN systems [1-6].

Currently, there is a great debate between the proponents of digital Time Division Multiple Access (TDMA) and Code Division Multiple Access (CDMA) (i.e., spread spectrum) as to which system provides maximum system capacity, without adding undue complexity, beyond that of today's analog systems. While it is clear that both TDMA and CDMA are very attractive relative to today's analog systems, we believe it may well be possible to realize a substantial *additional* gain in system capacity by the use of spatial diversity.

Towards this end, it is the purpose of this paper to set on sound theoretical footing some old ideas and proposals claiming that the capacity of most wireless systems can be significantly increased by exploiting the other dimension, space, that is available to the system designer. To capitalize on the spatial dimension, multiple antennas, spaced at least a half of a wavelength apart, are used to adaptively cancel the interference produced by users who are occupying the same frequency band and time slots. The interfering users can be in the same cell as the target user, and thus interference cancellation allows multiple users in the same bandwidth - in practice the number of users is limited by the number of antennas and the accuracy of the digital signal processors used at the receiver. The interferers can also be users in other cells (for frequency reuse in every cell), users in other radio systems, or even other types of radiating devices, and thus interference cancellation also allows radio systems to operate in high interference environments.

Optimum combining and signal processing with multiple antennas, is not a new idea [2-5]. But spurred on by new theoretical results, described in the sequel, it may be one whose time has come. Spatial diversity can be thought of as an overlay technique that can be applied to many wireless transmission systems, and it is known that with M antennas, $M-1$ interferers can be nulled out [3-5]. Consequently it can be used, perhaps in a proprietary manner, to increase the capacity of installed systems. Or, spatial diversity can be thought of as an alternative to the use of microcells to increase capacity. Microcells, while quite attractive, do create control (handoff) problems, require more base stations, and require sophisticated location planning for the new base stations. So, the added complexity of more antennas for optimum combining may be offset by the reduced complexity of the network controller, along with the reduction in the number of base stations, and the need for frequency planning (with frequency reuse in every cell). Furthermore, nulling of interferers can allow for low power transmitters to coexist with high power transmitters without a substantial decrease in performance and could lead to overlaid systems. Moreover, in some cases time-division retransmission [6] can be used to concentrate the complexity in the centralized base station (in this case the optimum combiner is used both as a receiver and transmitter array), so that the increased cost is amortized among all the users.

Use of spatial diversity is certainly made more compelling by the continued decrease in the cost of digital signal processing hardware, the advances in adaptive signal processing, and the above system benefits. Our continuous interest in this subject has recently yielded a new analytical result that is proven in the body of this paper: for a system with N users in a flat Rayleigh fading environment, optimum

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combining provided by a base station with $K+N$ antennas can null out $N-1$ interferers as well as achieve $K+1$ diversity improvement against multipath fading. Computer simulation shows that these results also hold with frequency-selective fading when optimum equalization is used at the receiver. In addition, the average error rate, or outage probability, behaves as if each user were either spatially or frequency isolated from the other users and derives the full benefit of the shared antennas for diversity improvement. These results provide a solid basis for assessing the improvement that can be achieved by antenna diversity with optimum combining.

In Section 2, we present theoretical results for flat fading and computer simulation results for frequency-selective fading with optimum combining. Experimental verification of interference suppression with flat fading is described in Section 3. In Sections 4 and 5, we discuss the application of optimum combining to the proposed North American standard for digital mobile radio, IS-54, and other systems, respectively. A summary is presented in Section 6.

II. Performance Analysis

A. System Description

Figure 1 shows a wireless system with N users, each with one antenna, communicating with a base station with M antennas. The channel transmission characteristics matrix $C(\omega)$ can be expressed as

$$C(\omega) = \left[C_1(\omega),\, C_2(\omega),\ldots,\, C_N(\omega) \right] \qquad (1)$$

where ω is the frequency in radians per second and the N column vectors (each with M-elements) $C_1(\omega)$, $C_2(\omega),\ldots,C_N(\omega)$ denote the transfer characteristics from the i^{th} user, $i = 1,2,\ldots,N$ to the j^{th}, $j = 1,2,\ldots,M$ receiver or antenna. Now consider the Hermitian matrix $C^\dagger(\omega)C(\omega)$, where the dagger sign stands for "conjugate transpose." If the vectors in (1) are linearly independent, for each ω, then the $N{\times}N$ matrix inverse, $(C^\dagger C)^{-1}$ exists. This is a mild mathematical requirement and will most often be satisfied in practice since it is assumed that users will be spatially separated.

At the receiver, the M receive signals are linearly combined to generate the output signals. We are interested in the performance of this system with the optimum linear combiner, which combines the received signals to minimize the mean-square error (MSE) in the output. An explicit expression was provided for the least obtainable total (for all N users) MSE in [7, Eq. (17)]. The formula for the minimum MSE for user "1" only, without loss of generality, is given by

$$(MSE)_{011} = \sigma_a^2 \frac{T}{2\pi} \int_{-\pi/T}^{\pi/T} \left[I + \frac{C^\dagger(\omega)C(\omega)}{N_0}\, \sigma_a^2 \right]_{11}^{-1} d\omega \qquad (2)$$

where $\sigma_a^2 = E\,|\,a_n^{(1)}\,|^2$, $[\quad]_{11}^{-1}$ stands for the "1 1" component of a matrix, T is the symbol duration, N_0 is the noise density, and $a_n^{(1)}$ are the 1^{st} user's complex data symbols.

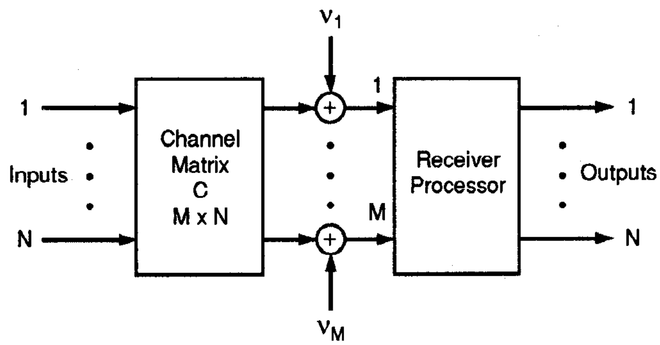

Fig. 1. Multiuser communication block diagram.

B. Flat Rayleigh Fading

With flat Rayleigh fading, the channel matrix $C(\omega)$ is independent of frequency and all the elements of C can be regarded as independent, zero-mean, complex Gaussian random variables with variance σ_i^2 for the i^{th} user, provided the antenna elements are sufficiently separated. This separation is typically about half a wavelength at the mobile because of local multipath and several wavelengths at the base station (because in many cases there is a line-of-sight from the base station to the vicinity of the mobile). Let us consider the high signal-to-noise case (which results in the "zero-forcing" optimum combiner solution). Under these assumptions (2) reduces to

$$(MSE)_{011} = (C^\dagger C)_{11}^{-1} N_0 \ . \qquad (3)$$

Since the minimum MSE for any signal-to-noise is always less than or equal to the MSE of the zero-forcing combiner (3), the zero-forcing solution serves as an upper bound on the MSE solution. For these reasons and the fact that it is easier to analyze the zero-forcing structure, we proceed in this paper with this approach. Using the MSE given by (3), we find that an exponentially tight upper bound on the conditional probability of error is given by [8, Eq. (16)]

$$P_{e1}(C) \le \exp\left\{ -\frac{\rho}{\sigma_a^2} \frac{1}{(C^\dagger C)_{11}^{-1}} \right\} , \qquad (4)$$

where ρ is the signal-to-noise ratio for user "1", i.e., $\rho = \dfrac{\sigma_a^2 \sigma_1^2}{N_0}$.

In order to analyze the performance of the general set-up, we must be able to determine the statistical properties of the random variable $\alpha = 1/(C^\dagger C)_{11}^{-1}$. From the definition of the inverse of a matrix we express this quantity as follows,

$$\alpha = \frac{\det\,(C^\dagger C)}{A_{11}} = \frac{\Delta_N(C_1,\ldots,C_N)}{\Delta_{N-1}(C_2,\ldots,C_N)} \qquad (5)$$

where $\det(\cdot)$ stands for determinant, A_{11} is the "11" cofactor, $\Delta_N(C_1,\ldots,C_N) = \det\,(C^\dagger C)$, and $\Delta_{N-1}(C_2,\ldots,C_N)$ is the determinant resulting from striking out the first row and first column of $C^\dagger C$. From the definition of the determinant

$$\Delta_N(C_1,C_2,...,C_N) = \sum \pm C_1^\dagger C_{i_1} C_2^\dagger C_{i_2} \cdots C_N^\dagger C_{i_N} \quad (6)$$

where the sum is extended over all $N!$ permutations of 1, 2,..., N, the "+" sign is assigned for an even permutation and "-" for an odd permutation, it can be seen that it is possible to factor out $C_1^\dagger$ on the left and C_1 on the right in each term. This factorization makes it possible to express Δ_N in the following form

$$\Delta_N(C_1,C_2,...,C_N) = C_1^\dagger F(C_2,C_3,...,C_N) C_1 \quad (7)$$

where F is an $M \times M$ matrix independent of C_1. By normalizing F by $\Delta_{N-1}(C_2...C_N)$ so that $F/\Delta_{N-1} = \tilde{M}$, we can express the quantity of interest as a positive quadratic form

$$\alpha = C_1^\dagger \tilde{M} C_1 \quad (8)$$

where $\tilde{M}$ is Hermitian and non-negative. Diagonalizing $\tilde{M}$ by a unitary transformation ϕ, we write for α

$$\alpha = C_1^\dagger \phi^\dagger \Lambda \phi C_1 = z^\dagger \Lambda z$$

$$= \sum_{i=1}^{M} \lambda_i \, |z_i|^2 \quad (9)$$

where Λ is $diag(\lambda_1 \cdots \lambda_M)$, λ_i's being the eigenvalues of $\tilde{M}$, $z = \phi C_1$, and $z_i = (\phi C_1)_i$, $i = 1, ..., M$.

Since C_1 is a complex Gaussian vector, so is z conditioned on ϕ. Also, the vectors C_1 and z possess identical statistics since ϕ is unitary. Therefore, conditioned on the eigenvalues, the random variable α is a weighted sum-of-squares of Gaussian random variables and therefore has a known probability distribution.

One would expect the actual distribution of α to be rather complicated since for example the characteristic function of α, conditioned on the eigenvalues, is readily evaluated in the form

$$E\left\{ e^{i\alpha\omega} \middle| \lambda_i, i=1,...,M \right\} = \prod_{i=1}^{M} (1 - 2\omega\lambda_i)^{-1} . \quad (10)$$

But since the eigenvalues are complicated nonlinear functions of the remaining $N-1$ vectors, $(C_2,C_3,...,C_N)$, the actual characteristic function of α, the average of (10) with respect to the eigenvalues, appears to be intractable. However, as shown in Appendix A, the eigenvalues of $\tilde{M}$ are equal to either 1 or zero, with $M-N+1$ eigenvalues equal to 1, and thus α is Chi-square distributed.

Applying this result in (4), we evaluate explicitly the average probability of error[1], i.e.,

[1] A more detailed derivation of the results in this section is presented in [9].

$$P_e = E_C P_e(C) \le E_C \exp\left\{ -\frac{\rho}{\sigma_a^2} \frac{1}{(C^\dagger C)^{-1}_{11}} \right\} = E_\alpha e^{-\frac{\rho}{\sigma_a^2}\alpha}$$

$$= E_z \exp\left\{ -\frac{\rho}{\sigma_a^2} \sum_{i=1}^{M-N+1} |z_i|^2 \right\} = \left[1 + \frac{\rho}{\sigma_a^2} \right]^{-(M-N+1)} . \quad (11)$$

Thus, the average probability of error with optimum combining, M antennas, and N interferers is the same as maximal ratio combining with $M-N+1$ antennas and no interferers.

The physical implications of this result are as follows. The error rate of a particular user is unaffected by all other users. It only depends on the user's own SNR, ρ. Of course, the price paid is in the diminished diversity benefits obtained for each user. For, when the number of antennas M equals the number of users N, the average error rate is as if there was only one antenna per user. But remarkably, the resulting performance is as if all the other users or interferers did not exist. The nulling-out of other users results only in reduced diversity benefits. But even when $M=N+1$, all users enjoy dual diversity, i.e., the addition of each antenna adds diversity to every user.

Furthermore, as stated previously, the above result is error rate performance with zero-forcing weights, whereby the interference is completely cancelled. In most practical systems, though, we don't need to cancel the interference, but only suppress it into the noise, and thus the minimum MSE combiner can achieve even better results than shown above. Note also that in most systems, the number of interferers is much greater than the number of antennas. However, these interferers are usually much weaker than the desired signal (rather than equal to it, as we have considered), and optimum combining can still achieve gains over maximal ratio combining (see, e.g., [11]), although our theoretical results (11) no longer apply.

C. Frequency-Selective Fading

With frequency-selective fading, unfortunately, no closed form analytical results exist as for the flat fading case. The problem is complicated since in this case the variances of the output noise samples are complicated functionals of the matrix channel characteristics, $C(\omega)$. The performance of optimum (MSE) combining and optimum equalization (linear equalization with an infinite length tapped delay line) has been previously studied by computer simulation in [17] for cochannel interference and frequency-selective fading. These results showed that the performance improves with frequency-selective fading and optimum equalization. Here we want to verify that for the zero-forcing combiner and optimum equalization, the capacity and performance gains we obtained with flat fading still hold or are even improved.

With frequency-selective fading and optimum equalization, the MSE is given by (2). For the zero-forcing combiner, with sufficiently high signal-to-noise ratio, in (2) the I matrix is negligible as compared to $\frac{C^\dagger(\omega)C(\omega)}{N_o}\sigma_a^2$ and is dropped.

Using the probability of error bound for given MSE of [8, Eq. (16)] (as in Section 2.2), we obtain an exponentially tight bound on the conditional probability of error given by

$$P_e(C(\omega)) \le \exp\left\{-\frac{\rho}{\sigma_a^2}\frac{1}{\sigma^2(C)}\right\} \tag{12}$$

where

$$\sigma^2(C(\omega)) = \frac{T}{2\pi}\int_{-\frac{\pi}{T}}^{\frac{\pi}{T}} \left[C^{\dagger}(\omega)C(\omega)\right]^{-1}_{11} d\omega \ . \tag{13}$$

The outage probability as well as the average probability of error depends in a complicated way on the statistical characterization of the matrix $C(\omega)$.

If we assume that the propagation mode is by uniformly distributed scatterers and delay spread cannot be neglected, then a reasonable statistical model for $C(\omega)$ is the following. For each frequency ω, every entry in $C(\omega)$ is complex Gaussian, but at different frequencies the entries are correlated. Specifying the multidimensional correlation function provides a complete statistical characterization of the matrix medium. For this model, which is often referred to as the "frequency-selective fading" Rayleigh medium, we can derive an upper bound on the average probability of error. Also, for a two ray model of the frequency-selective Rayleigh process for each entry of the matrix C, we have carried out Monte Carlo evaluations. We will discuss these results later, but first we provide an outline of our bounding technique.

Note that from the properties of the matrix $C^{\dagger}(\omega)C(\omega)$, irrespective of the statistics, we can always express the noise variance as

$$\sigma^2(C(\omega)) = \left\langle \frac{1}{\sum_{l=1}^{M-N+1} |z_l(\omega)|^2} \right\rangle_{\omega} \tag{14}$$

where $\langle\cdot\rangle_\omega = \frac{T}{2\pi}\int_{-\frac{\pi}{T}}^{\frac{\pi}{T}} \left[\cdot\right] d\omega$ and $z_i(\omega) = \phi_i^{\dagger}(\omega)C_1(\omega)$ where $\phi_i(\omega)$ are the eigenvectors of the matrix $\tilde{M}$. We now note that for each frequency

$$\alpha(\omega) = \sum_{i=1}^{M-N+1} |z_i(\omega)|^2 \tag{15}$$

is Gamma distributed with probability density

$$p(\alpha) = \frac{\alpha^{K-1}e^{-\alpha}}{(K-1)!} \tag{16}$$

where $K=M-N+1$.

Making use of these facts, an upper bound on the average probability of error is given by (see Appendix B)

$$\overline{P_e} = E_C P_e(C(\omega)) \le d_{M-N}\left[\frac{\sigma_a^2}{\rho}\right]^{M-N} \tag{17}$$

where $d_{M-N} = \frac{1\cdot3\cdot5\cdots[2(M-N)-1]}{(M-N)!}$. While this may appear to be a loose upper bound, it does indicate that when the number of antenna elements is not much greater than the number of users or interferers we only lose the diversity benefit from one additional antenna.

As an illustration, suppose that $M-N=1$, i.e., one more antenna element than users. Our bound indicates that $\overline{P_e} \le 1/\rho$ for a binary system when $\sigma_a^2=1$. On the other hand, when only flat fading is present, we can expect $\overline{P_e} \le \frac{1}{\rho^2}$.

In actual Monte Carlo simulation and evaluation of averages presented below, we found that the average error rates were much lower than predicted from (17).

Before proceeding, we note that with a two ray model of frequency-selective fading with $N=1$ (no interference), [2] provides bounds showing that the average bit error rate decreases with increasing time delay between the two multipath rays when optimum combining and equalization is used. For this two ray model, the ijth element of $C(\omega)$ is given by

$$c_{ij}(\omega) = a_{ij} + b_{ij}e^{-j\omega\tau_i} \tag{18}$$

where a_{ij} and b_{ij} are complex Gaussian random variables with zero mean and variance 1/2, and τ_i is the time delay between the two rays.

To gain insight into the behavior of average error rate versus delay spread, we used Monte Carlo simulation to derive 1000 channel matrices C and numerically calculated the average bit error rate for each channel from (12). The entries in C are given in (18). The bit error rate averaged over these 1000 C matrices is shown in Figures 2 and 3 for ρ/σ_a^2 = 18 dB. Figure 2 shows the average bit error rate versus τ/T, where T is the symbol duration, for $M=N$ with a) frequency-selective fading of the desired and interfering signals, $\tau_1=\tau_2=\cdots=\tau_N=\tau$, where τ_1 is the time delay between the two multipath rays of the desired signal and $\tau_2,\cdots,\tau_N$ is the time delay of the interfering signals, b) frequency-selective fading of the interferers only, $\tau_1=0$, $\tau_2=\cdots=\tau_N=\tau$, and c) frequency-selective fading of the desired signal only, $\tau_1=\tau$, $\tau_2=\cdots=\tau_N=0$. At $\tau/T=0$, the simulation results for $M=1$, 2, and 3 should be equal to the theory (flat-fading) point shown. The simulation results differ from the theory because only 1000 samples were used due to CPU time limitations. For all three values of M, the bit error rate (BER) decreases with τ/T until $\tau/T=1$, and then remains approximately constant, because the signals in the two rays are uncorrelated in this case since the bandwidth of the signal is equal to the data rate. At $\tau/T=1$, the simulation results for a) $\tau_1=\tau_2=\cdots=\tau_N=\tau$ are in agreement with theoretical results [2] for a single signal with optimum equalization, within the sampling error, with slightly degraded performance for cases b) and c). Thus, with

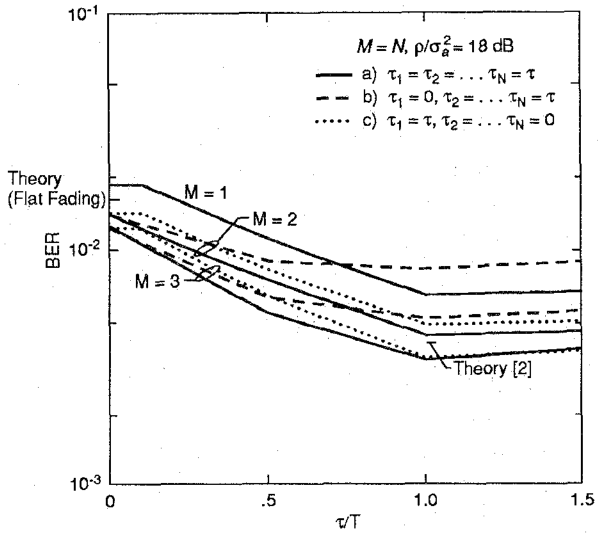

Fig. 2. Effect of frequency-selective fading for $M=N$, with optimum combining and equalization.

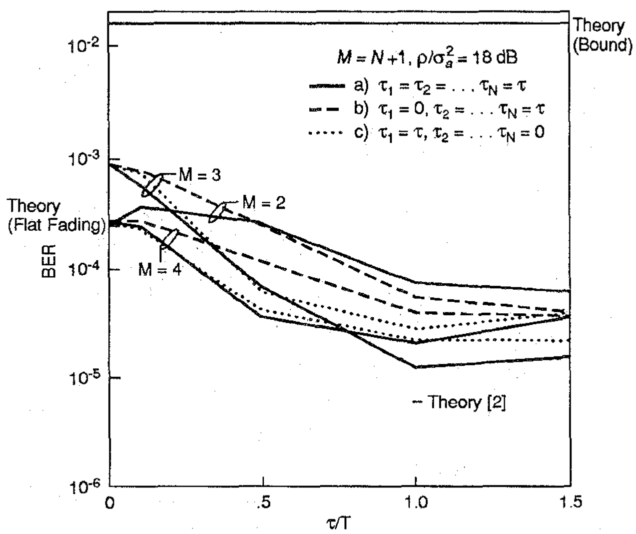

Fig. 3. Effect of frequency-selective fading for $M=N+1$, with optimum combining and equalization.

frequency-selective fading and optimum combining and equalization with M antennas, each of the $N=M$ users have the same performance as that of a single antenna system without interference.

Figure 3 shows similar results for $M=N+1$. Thus, with $M=N+1$, we can achieve dual diversity for all users, with each user having the performance gain of a single antenna/user system with frequency-selective fading. Therefore, our simulation results show that our results for flat fading also hold with frequency-selective fading and optimum equalization.

III. Experimental Results

To demonstrate and test the interference nulling ability of optimum combining in a fading environment, an experimental system was built. Figure 4 shows a block diagram of the experiment, which consisted of 3 users, a 24 channel Rayleigh fading simulator, 8 receive antennas, and a DSP32C processor at the receiver. The three remotes' signals used QPSK modulation, at a common 50 MHz IF frequency, consisting of a biphase data signal and a quadrature biphase signal with a pseudorandom code that was unique to each user. This pseudorandom code was used to generate the reference signal at the receiver (see [3,4]) that was used to distinguish the users. Thus half of the transmitted signal energy was allocated for reference signal generation only. The fading simulator generated the 8 output signals for the antennas by combining the three remotes' signals with independent flat, Rayleigh fading between each input and antenna output. The fading rate of the simulator was adjustable up to 81 Hz. The outputs of the simulator were demodulated by the 8 antenna subsystems, A/D converted, multiplexed, and input to a DSP32C. This DSP32C used the Least Mean Square (LMS) algorithm ([10], see also [3,4]) to acquire and track one of the remote's signals. With our program in the DSP32C, the maximum weight update rate was 2 kHz, and the data rate was set to 2 kbps for convenience (although any data rate greater than 2 kbps could have been used). The step size of the LMS algorithm was limited to keep the change in weights small enough so that the data was not significantly distorted by the weights and so that the algorithm remained stable.

Experimental results were obtained for the case of equal (averaged over the Rayleigh fading) received-power signals, as in the theoretical results of the previous section. Thus, with two interferers, the desired-signal-to-interference-power ratio was -3 dB and the BER without optimum combining was approximately 0.5. Our experimental results showed that optimum combining reduced the BER below 10^{-2} (suitable for mobile radio) even with a fading rate of 81 Hz. Note that this corresponds to a data rate (2 kbps) to fading rate ratio of 25, which is much faster fading than in most wireless systems (see Sections 4 and 5). Thus, the experiment successfully demonstrated that 2 interferers with power equal to the desired signal can be suppressed for a 3-fold capacity increase (i.e., 3 users in one channel) in a fast fading environment. Noise on the circuitry backplane limited the accuracy of the A/D to 6 bits, which did not allow verification of the 6-fold diversity improvement predicted by (11) for M=8 and N=3, or precise calculation of the level of interference suppression.

IV. Application to IS-54

To illustrate the application of adaptive antennas to proposed wireless communication systems, in this section we consider the proposed North American standard for digital mobile radio, IS-54. In this cellular TDMA system, 3 remotes communicate with the base station in each 30 kHz channel within a 824 to 849 MHz (mobile to base) and 869-894 MHz (base to mobile) frequency range, at a data rate of 13 kbps per user using DQPSK modulation. Each user's slot contains

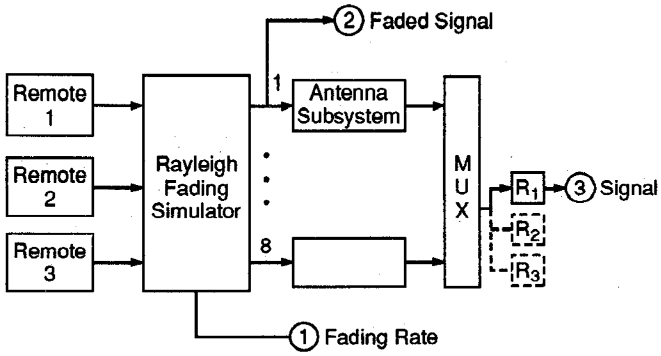

Fig. 4. Experimental system.

324 bits, including a 28 bit synchronization sequence, 12 bit user identification sequence, plus 260 data bits, resulting in a data rate for each channel of 48.6 kbps (24.3 kbaud)[1].

In small cells in urban areas, the multipath delay spread is usually a fraction of a symbol, and equalization is not needed. However, in larger cells, e.g., in suburban or rural areas, the delay spread may be as large as a symbol and, in this frequency-selective fading environment, equalizers may be used at both the base station and mobiles. Base stations use two antennas for reception with selection or postdetection combining (see below), while two antennas are only an option at the mobile. This is due to the fact that mobile to base station transmission requires greater improvement than the reverse link, because portable phones, which transmit less power, must be accommodated along with the phones in vehicles. Thus, today there are two classes of mobile transmitters: portables and mobile units, as well as two classes of mobile receivers: those with and without diversity. A frequency reuse factor of 7 is generally used in order to provide adequate service for all classes of transmitters and receivers, which means that for each user there are up to 6 cochannel interferers two cells away.

The IS-54 application of optimum combining differs from the systems studied in Sections 2 and 3 in that, typically, the number of interferers is greater than the number of antennas, but the interferers have lower power than the desired signal. Thus, in IS-54 optimum combining cannot completely null all interferers, but can decrease their power in the array output by a few dB. This can suppress interference below the noise level and decrease the required receive signal-to-interference-plus-noise ratio (SINR), which permits lower frequency reuse factors and, thus, higher capacity (as shown below).

A. Mobile To Base

Let us next consider how adaptive antennas can be used to improve the performance and increase the capacity of this system. We will first consider the weaker link from the mobile to the base station. Figure 5 shows a block diagram of the system to be considered. At the base station there are multiple antennas, but only one antenna at each mobile (multiple mobile antennas will be considered later). The antennas are positioned such that the fading of each signal at each antenna is independent (see Section 2.2). At the base station, the received signals are linearly combined to reduce the effects of multipath fading and eliminate interference from other users. Figure 6 shows a block diagram of the M element adaptive array. The signal received by the ith antenna element is passed through a tapped delay line equalizer (only two taps are shown in the figure for simplicity) with controllable weights. The weighted signals are then summed to form the array output. Note that the tapped delay line equalizer is required only in areas with large delay spread. In congested urban areas, there would be only one tap per antenna.

The weights can be calculated by a number of techniques. Here, we will consider two techniques, the LMS algorithm (as in Section 3) [10] and Direct Matrix Inversion (DMI)[10].

With DMI, the weights are given by

$$\mathbf{w} = \hat{\mathbf{R}}_{xx}^{-1}\hat{\mathbf{r}}_{xd} \ , \tag{19}$$

where

$$\mathbf{w} = [w_{11} \cdots w_{ML}]^T \ , \tag{20}$$

w_{ij} is the weight for the jth tap on the i antenna element, the superscript T denotes transpose, L is the number of taps in each equalizer, M is the number of antennas, the receive signal cross-correlation matrix is

$$\hat{\mathbf{R}}_{xx} = 1/K\sum_{j=1}^{K}\mathbf{x}(j)\mathbf{x}^{\dagger}(j) \ , \tag{21}$$

K is the number of samples used,

$$\mathbf{x} = [x_1(j)x_1(j-1)\cdots x_1(j-L+1)\cdots x_M(j-L+1)]^T \ , \tag{22}$$

$x_i(l)$ is the received signal at antenna i in the lth bit interval, the reference signal correlation vector is

$$\hat{\mathbf{r}}_{xd} = 1/K\sum_{j=1}^{K}\mathbf{x}(j)r^*(j) \ , \tag{23}$$

the superscript * denotes complex conjugate, and $r(j)$ is the reference signal. The reference signal is used by the array to distinguish between the desired and interfering signals at the receiver. It must be correlated with the desired signal and uncorrelated with any interference. The generation of the reference signal is discussed below.

The LMS algorithm has lower computational complexity than DMI and its complexity increases linearly with the number of taps, while DMI's complexity increases much faster than linearly because the technique uses matrix inversion (19). The LMS algorithm converges to the optimum weights at a slower rate than DMI, however, and its convergence speed depends on the eigenvalues of $\hat{\mathbf{R}}_{xx}$, i.e., the power of the desired signal and interferers. Thus, weak interferers are tracked at a much slower rate than the desired signal. Although this was not a problem in the experiment of Section 3, where the desired signal and interference had the same power, it is a serious problem in IS-54. As shown in [11], the LMS algorithm cannot track weak interference in IS-54. DMI, however, has a convergence speed that is independent of the signal powers and can converge (with less

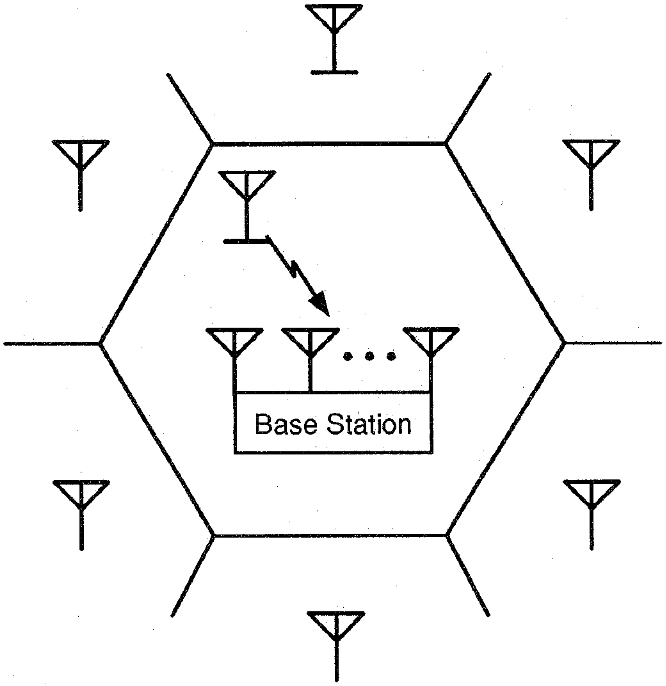

Fig. 5. Cellular radio system with multiple antennas at the base station.

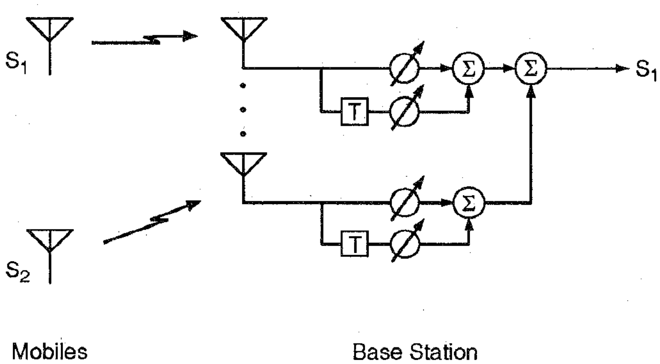

Fig. 6. Block diagram of an M element adaptive array.

than a 3 dB SNR degradation) to the optimum weights with only $2ML$ samples (K in (21)) [10, p. 297]. In [11], we showed that with 2 antennas and 1 interferer, DMI can acquire and track both the desired signal and interferer in IS-54, with the performance of optimum combining within 1 dB of the predicted ideal tracking performance. Thus, we will only consider DMI for IS-54.

Next, consider the reference signal generation. For weight acquisition, we will use the known 28 bit synchronization sequence as the reference signal, using DMI to determine the initial weights. After weight acquisition, the output signal consists mainly of the desired signal and (during proper operation) the data is detected with a BER that is not more than 10^{-2} to 10^{-1}. Thus, we can use the detected data as the reference signal. Note that since there will be processing delay in determining the data (generally, a few msec, which is not perceptible to the user), DMI must use delayed samples of the received signal and array output.

For IS-54, the fading rate can be as high as 100 Hz (for 75 mph (120 km/hr) vehicles at 900 MHz), and thus with 24.3 kbaud the channel can completely change in as little as 243 symbols. This is slow enough so that the channel does not change significantly over the window of symbols, K, used by DMI for acquisition (K=14) and tracking (K=14 was used in [11]).

Another issue is how to distinguish the desired user's signal from other users in other cells. In IS-54, the synchronization sequence for a given time slot is the same for all users, but is different for each of the six time slots in each frame. Since base stations operate asynchronously, signals from other cells have a high probability of having different timing (since there are 972 symbols per frame) and being uncorrelated with the reference signal for the desired signal. The 12 bit user identification code can be used to verify that the correct user's signal has been acquired.

Let us now consider the improvement in the performance and capacity of IS-54 with optimum combining. In congested urban areas with small delay spread, the performance improvement with optimum combining can be calculated from previous papers. When equalization is not required, differential detection of the DQPSK signal can be used, followed by postdetection combining of the signals from the two antennas. Theoretically, such a system requires a 17.2 dB average SINR to operate at an average BER of 10^{-3}, although slightly higher BER's (and lower SINR) may be acceptable. Maximum ratio combining (which maximizes the signal-to-noise ratio) decreases this by 1.6 dB [12], while optimum combining decreases the required average SINR by 3.6 dB [3][2]. Three antennas with optimum combining permit a 9.2 dB reduction, which increases to 12.4 dB with 4 antennas and 15 dB with 5 antennas [3]. Thus, with optimum combining at the base station only, the required portable transmit power can be reduced by these amounts, allowing for good performance with very low power portables.

In terms of frequency reuse, the average SINR is 18.7 dB with a frequency reuse factor of 7, while the worst case SINR (which occurs when the desired mobile is at the edge of the cell and the interfering mobiles are all as close as possible to the desired mobile's base station) is 14.4 dB [13]. A frequency reuse factor of 4 lowers the SINR to 13.8 (average) and 7.9 dB (worst case), which decreases to 11.3 and 4.3 dB with a frequency reuse factor of 3 and 1.8 and -7.8 dB with frequency reuse in every cell. Thus, with optimum combining at the base station with two antennas, a frequency reuse factor of 4 may be possible except in some worst cases. However, we may be able to avoid the worst cases by using dynamic channel assignment, whereby, the channel of a user is changed if the interference is too strong, i.e., if the adaptive array cannot adequately suppress the interference. Although the interference suppression, and thus the BER, varies at the fading rate, channel reassignment need only be based on longer term performance. For example, if the desired mobile is near the cell boundary and all the mobiles using the same

[2] Note that optimum combining reduces the signal to noise ratio as compared to maximum ratio combining in order to reduce the interference power and achieve a lower SINR.

channel are close to the desired mobile's base station, frequent high BER's would result in the desired mobile being assigned to a channel where the interfering mobiles are farther away. Note that the range between the average and worst cases increases as the frequency reuse factor decreases, and thus dynamic channel assignment becomes more useful. Dynamic channel assignment may even permit a frequency reuse factor of 3, which more than doubles the capacity of the mobile to base link (as compared to a reuse factor of 7), while 4 or 5 antennas may be required for frequency reuse in every cell. Alternatively, we may wish to increase the capacity of only a few congested cells, keeping the frequency reuse factor of 7. In this case we can achieve an M-fold capacity increase for the mobile to base station link by adding M antennas with optimum combining to the base station. However, with frequency reuse in a cell which uses the same base station, we would need to either time offset the multiple users in the same channel or modify the IS-54 standard to allow for different synchronization sequences for the same time slot with multiple users.

In areas with large delay spread, the performance improvement with optimum combining and equalization can only be estimated from our results. As shown in Section 2.3, the performance of optimum combining generally improves with frequency-selective fading (when optimum equalization is also used) and therefore our above results for the flat fading IS-54 system should also hold with frequency-selective fading.

B. Base To Mobile

Because the base to mobile link is the stronger link when low power portables are accommodated, the required number of antennas and complexity of the mobiles will generally be lower than that of the base station. Indeed, low power portables can be accommodated without any changes at the mobiles. For higher frequency reuse factors, though, more than one antenna might be necessary at the mobile. From the previous section, two antennas with optimum combining at the mobile may be required for a frequency reuse factor of 3. Many mobiles currently have 2 antennas (with selection diversity), though, and dual diversity is available with a single antenna. Unfortunately, frequency reuse in every cell might require 4 or 5 antennas at the mobile, which is currently not commercially practical. However, there are techniques whereby multiple transmit antennas at the base station can be used in such a way that only one or two mobile antennas are needed. One technique is transmit diversity [18] which is easiest to implement in systems with small delay spread (e.g., in urban areas). Another technique, perhaps more useful in rural or suburban areas, is as follows.

The above results assume omnidirectional transmission from one base station antenna. Further reductions in the required number of antennas and the complexity at the mobiles may be possible in cells where a line-of-sight exists between the base station antennas and the vicinity of the mobile. This is often the case (especially in suburban and rural areas) because the base station antennas are located fairly high off the ground. In these cases, multiple antennas at the base station can be used to transmit with the same antenna pattern as received (using weights that are the complex conjugate of the received weights), as shown in Figure 7. With line-of-sight to the mobile vicinity, this results in increased gain in the direction of the mobile and decreased gain in the direction of any interferers.[3] Thus, cochannel interference from a base station to mobiles in other cells is reduced, and each mobile receives its own signal with very little cochannel interference. Although further studies are needed, this could result in a single antenna at the mobile being adequate for a frequency reuse factor of 3, and only 2 antennas at the mobile required for frequency reuse in every cell.

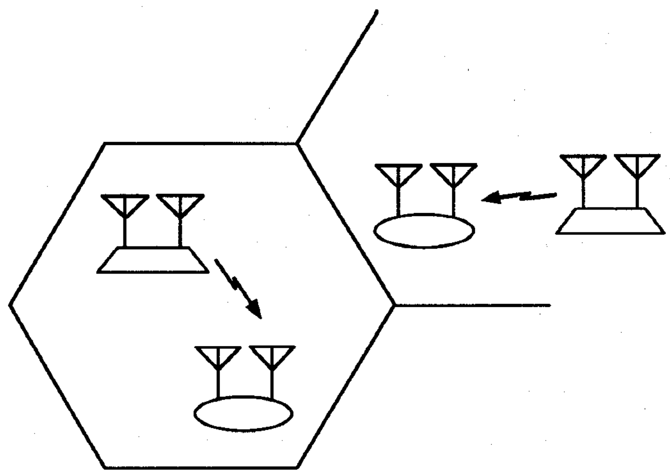

Fig. 7. Directional transmission from the base station.

C. Other Issues

Issues for IS-54 requiring further study include the following. First, DMI should have $K \geq 2ML$ (see (21)) for accurate weight calculation. Thus, with the 14 symbol synchronization sequence, ML should be less than 7, which places a limit on the capacity growth with IS-54. Second, with DSP implementation of optimum combining, the DSP bit accuracy limits the allowable dynamic range of the signals, and thus the maximum power of an interferer that can be suppressed. Some power control may therefore be required, although this should be a much less stringent requirement than in current cellular systems. Third, our results assume independent Rayleigh fading at each antenna, which may not always hold. In [14], we studied the effect of correlated fading and showed that adaptive array performance was not significantly degraded with correlation as high as 0.5. Note that if widely-spaced antenna elements (macrodiversity) are used, the average received signal level can vary among the antennas, and this may degrade performance. Since even without multipath, $M-1$ interferers can be eliminated with M antennas, optimum combining looks promising in a wide variety of environments, but further study is needed. Fourth, as shown in [3], further improvement can be obtained by using sectorized antennas. By using antennas with 120° beamwidths, the number of interferers is reduced by a factor of

[3] With multipath around the base station, this technique may not work because the transmit and receive frequencies may differ by more than the correlation bandwidth, resulting in nonreciprocal transmit and receive channels.

3, and, for frequency reuse in every cell, only 5 antennas are required in the worst case scenario [3]. For typical scenarios (using dynamic channel assignment), fewer antennas would be required. Additional sectorization to 60° beamwidths further reduces the number of interferers and thus the required number of antennas. Thus, with a few more antennas, channels can be reused in a cell as these interferers can be suppressed by the adaptive array. The capacity can therefore be increased from 7-fold to 14-fold, etc., by adding additional antennas. Note that this increase in capacity occurs without increasing the number of base stations, and thus the number of handoffs does not increase (unless sectorized antennas are used), allowing graceful growth to very high capacity. Fifth, for the frequency-selective fading environment, we need to determine the performance of DSP implementations of the techniques for joint optimum combining and equalization. Techniques that jointly detect all the signals (not just the desired signal) also need to be proposed and studied.

Finally, let us consider how IS-54 could be modified to allow for better performance and higher capacity. With the current system, the base stations operate asynchronously, which results in interferers turning on and off during each time slot. If the interference power is too high, during tracking the output data bits may become the interferers' data, and DMI may acquire and track the interferer and null the desired signal. This problem can be solved by operating all the base stations synchronously, such that the interferers change slowly (at the fading rate) during the time slot. With synchronized base stations, though, each user in a given time slot would require a unique code so that the reference signal is correlated only with that signal. In addition, these codes should have low cross-correlation for better weight acquisition by DMI (19). The length of the synchronization sequences could also be increased to allow a greater number of antennas to be used, resulting in greater capacity through higher frequency reuse.

In summary, in IS-54, optimum combining at the base station with two antennas may allow more than a doubling of capacity and accommodate lower power portables. Additional antennas at the base station (perhaps with two antennas at each mobile) could potentially permit frequency reuse in every cell. Thus, adding antennas allows for graceful capacity growth and performance improvement.

V. Other Applications

Optimum combining is an overlay technique that can be used to improve the performance of many interference-limited systems. Below, we consider issues for possible applications with the mobile radio systems of CDMA and GSM, and with slower fading systems, such as indoor radio.

A. CDMA

In QUALCOMM's CDMA mobile radio system, the number of interferers for each user is very large, on the order of 100 or more. Thus, for practical systems, the number of antennas is far less and our results (11) are not useful. Furthermore, optimum combining is not as effective against the interference as in IS-54. Since the CDMA system uses a RAKE [15] receiver to separate the multipaths, however, the number of interferers is on the order of the product of the number of antennas times the number of RAKE taps and interference can be suppressed. In the CDMA system, the weak link is the mobile to the base station, since the different multipaths from each mobile to the base station decrease the orthogonality of the interference, which is maintained with base station to mobile transmission. Thus, the complexity of optimum combining needs only be added to the base station. Furthermore, since all signals are detected at the base station, knowledge of the interfering signals can be used to further improve performance [16] (this can also be done in IS-54). Note that in systems with a line-of-sight from the base station to the vicinity of the mobile, with optimum combining the base station can generate a main beam in the direction of the desired mobile that reduces interference from other directions, similar to using sectorized antennas (without the need for handoffs between sectors).

B. GSM

Optimum combining can also be used in GSM [1] to improve performance and increase capacity. In this system, a training sequence (26 bits) is available in the middle of each time slot for weight acquisition, as in IS-54. However, as compared to IS-54, the signaling rate is higher (270.833 kbps) and the time slots are shorter (577 μsec), such that the weights do not need to be adapted within each time slot after acquisition. Optimum combining may also help in reducing intersymbol interference, particularly since the multipath delay can be more than four bits.

C. Slow Fading Systems

For stationary or portable radio systems, the fading rate will be much lower than for mobile radio systems, which permits better suppression of interference. In these systems, optimum combining can suppress interference from other devices which can make the system more robust and allow reliable operation of the system in the ISM and other frequency bands where these systems may operate as secondary users. In addition, optimum combining can also suppress signals from users of other systems, which can permit the system to operate even with similar radio systems (at the same frequency) in the same building or area.

Slower fading rates, relative to the data rate, can also makes adaptive retransmission with time division possible. With adaptive retransmission, the base station transmits at the same frequency as it receives, using the complex conjugate of the receiving weights. With time division, a single channel is time shared by both directions of transmission. Thus, during mobile-to-base transmission the antenna element weights are adjusted to maximize the SINR at the receiver output. During base-to-mobile transmission, the complex conjugate of the receiving weights are used so that the signals from the base station antennas combine to enhance reception of the signal at the desired mobile and to reduce the power of this signal at other mobiles. Therefore, the improvement in the performance with optimum combining as compared to maximal ratio combining should be similar to that at the base stations. Thus, both the mobile and the base station receivers benefit from optimum combining with the complexity of optimum combining and multiple antennas at the base station only. The actual improvement for a given mobile depends on the interference environment of every base station, however. Note that for this technique to be effective, the channel must not change significantly during retransmission. In addition, this technique works properly only when all signals use adaptive retransmission with *synchronized* time division. This includes interference from other systems (e.g., multiple systems within a building). If this is not the case (as with interference from other RF equipment), then optimum combining must also be used at the remote.

In slow fading systems, the weights can be adapted using a preamble for the reference signal, and then the weights can be frozen for a short period of time. Thus, the preamble may use only a small fraction of the total number of bits. However, if asynchronous interference is present, then a code, unique to each user, must be continuously transmitted. One technique, which was considered previously for the LMS algorithm ([3,4] and used in the experiment described in Section 3), is to use spread spectrum techniques. Specifically, we add a different pseudonoise code to each signal in phase quadrature to the original signal. Thus, a known signal is always available for reference signal generation, and the detected data is not used to distinguish users.

VI. Conclusions

For a broad class of interference-dominated wireless systems including mobile, personal communications, and wireless PBX/LAN networks, we have shown that a significant increase in system capacity can be achieved by the use of spatial diversity (multiple antennas) and optimum combining. In many systems, time-division retransmission can be used so that the processing complexity can be shared at the base station. A significant increase in user capacity may be achieved with a modest increase in complexity. Moreover, the system naturally lends itself to modular growth and improved performance by increasing the number of antennas. A true assessment of the value of antenna diversity can only be attained by a feasibility experiment with target system parameters in an actual operating environment.

Appendix A

THEOREM

Let $C_1, C_2, ..., C_N$ be N linearly independent complex M-vectors, $M \geq N$. Form the $N \times N$ non-negative Hermitian matrix $H = C^\dagger C$ with entries $H_{ij} = C_i^\dagger C_j$ $i,\ j = 1,2,...,N$. Let $\Delta_N\ (C_1, C_2, ..., C_N) = \det(C^\dagger C)$ and $\Delta_{N-1}(C_1, ..., C_{l-1}, C_{l+1}, ..., C_N)$ be the determinant of $C^\dagger C$ by striking out the l^{th} row and l^{th} column of $C^\dagger C$. From the definition of the determinant, the ratio Δ_N / Δ_{N-1} is expressed as a quadratic form in the components of C_l,

$$\Delta_N / \Delta_{N-1} = C_l^\dagger \tilde{M} C_l \tag{A.1}$$

where $\tilde{M}$ is non-negative Hermitian.

The matrix $\tilde{M}$ has $N-1$ eigenvalues equal to zero and $M - N + 1$ eigenvalues equal 1.

We have discovered and proved this unexpected result by examining the detailed structure of the matrix $\tilde{M}$. We thank J. E. Mazo for supplying a concise proof. Below, we supply the spirit of Mazo's proof rather than our original and lengthy proof.

PROOF

In (A.1) set C_k in place of C_l, $k=1,2, \cdots, N$, $k \neq l$ and obtain

$$\frac{\Delta_N(C_1, C_2, \ldots, C_{l-1}, C_k, C_{l+1}, \ldots, C_N)}{\Delta_{N-1}(C_1, C_2, \ldots, C_{l-1}, C_{l+1}, \ldots, C_N)} = C_k \tilde{M} C_k \tag{A.2}$$

$$= 0, \quad \text{all } k \neq l \ .$$

The last equality holds since Δ_N now has two identical columns. Since $\tilde{M} \geq 0$ and $C_k^\dagger \tilde{M} C_k = 0$ holds for $(N-1)$ linearly independent C_k's by hypothesis, $(N-1)$ eigenvalues of $\tilde{M}$ must be equal to zero. This establishes the first claim.

To establish the second claim, consider a vector y which is in the orthogonal subspace spanned by $(C_1, C_2, \cdots, C_{l-1}, \ , C_{l+1}, \cdots, C_N)$. Now in (A.1) replace C_l by y and since y is orthogonal to all the other C's we have

$$\frac{\Delta_N(C_1, \ldots, C_{l-1}, y, C_{l+1}, \ldots, C_N)}{\Delta_{N-1}(C_1, \ldots, C_{l-1}, C_{l+1}, \ldots, C_N)} = y^\dagger y \tag{A.3}$$

$$= y^\dagger \tilde{M} y \ .$$

There are exactly $M-N+1$ such vectors in the orthogonal subspace and therefore each y_n must satisfy

$$y_n^\dagger \tilde{M} y_n = y_n^\dagger y_n \ , \ n = 1,2,...,M-N+1 \ , \tag{A.4}$$

which implies that there are as many eigenvalues $(M-N+1)$ equal to unity.

Appendix B

Let the noise (real or imaginary portion) in the received signal at the detector be denoted as q with variance given by (14) for given $C(\omega)$. Then the average error probability is given by

$$\overline{P_e} = Pr\left[q \geq 1\right] = \int_1^\infty p(q)dq \leq \int_1^\infty q^{2n} p(q) dq \leq Eq^{2n}$$

$$= E_{\alpha(\omega)} E\left[q^{2n} \mid \alpha(\omega)\right] \ , \tag{B.1}$$

for some n such that the integral exists. But since $q^{2n} \mid \alpha(\omega)$ is Gaussian, we can express its $(2n)^{th}$ moment and obtain,

$$\overline{P_e} \leq E_{\alpha(\omega)} \left[\frac{\sigma_a^2}{\rho} \left\langle \frac{1}{\alpha(\omega)} \right\rangle \right]^n 1 \cdot 3 \cdot 5 \cdots (2n-1) \ . \tag{B.2}$$

Using Jensen's inequality, we upper bound (B.2) by,

$$\overline{P_e} \leq \left[\frac{\sigma_a^2}{\rho}\right]^n \left\langle E_{\alpha(\omega)} \left[\frac{1}{\alpha^n(\omega)}\right] \right\rangle 1 \cdot 3 \cdot 5 \cdots (2n-1) \ . \tag{B.3}$$

Evaluating the inner average using the Gamma distribution we get,

$$E_{\alpha(\omega)} \frac{1}{\alpha^n(\omega)} = \frac{1}{(K-1)!} \int_0^\infty \alpha^{K-1-n} e^{-\alpha} d\alpha = \frac{1}{(K-1)!} \ . \tag{B.4}$$

We choose the largest $n=K-1$ so as to make the bound as tight as possible. Therefore the sought-after upper bound is as given in (17).

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