Cut-off Rate and Signal Design for the Quasi-static Rayleigh Fading Space-Time Channel[1]

Alfred O. Hero and Thomas L. Marzetta

March 29, 2001

Submitted to *IEEE Transactions on Information Theory*, March 2000 - revised Dec. 2000

Abstract

We consider the computational cut-off rate and its implications on signal design for the complex quasi-static Rayleigh flat fading spatio-temporal channel under a peak power constraint where neither transmitter nor receiver know the channel matrix. The cut-off rate has an integral representation which is an increasing function of the distance between pairs of complex signal matrices. When the analysis is restricted to finite dimensional sets of signals interesting characterizations of the optimal rate-achieving signal constellation can be obtained. For arbitrary finite dimension, the rate-optimal constellation must admit an equalizer distribution, i.e., a positive set of signal probabilities which equalizes the average distance between signal matrices in the constellation. When the number N of receive antennas is large the distance-optimal constellation is nearly rate-optimal. When the number of matrices in the constellation is less than the ratio of the number of time samples to the number of transmit antennas, the rate-optimal cut-off rate attaining constellation is a set of equiprobable mutually-orthogonal unitary matrices. When the SNR is below a specified threshold the matrices in the constellation are rank one and the cut-off rate is achieved by applying all transmit power to a single antenna and using orthogonal signaling. Finally, we derive recursive necessary conditions and sufficient conditions for a constellation to lie in the feasible set.

Keywords: space-time coding, flat Rayleigh fading, spatial diversity, optimal signal design, unitary constellations, minimal distance codes, peak power constraints.

Corresponding author:

Professor Alfred O. Hero III
Room 4229, Dept. of EECS
University of Michigan
1301 Beal Avenue
Ann Arbor, MI 48109-2122.
Tel: (734) 763-0564.
Fax: (734) 763-8041.
e-mail: hero@eecs.umich.edu.

[1]A. Hero (corresponding author) is with the Dept. of Electrical Engineering and Computer Science, The University of Michigan, Ann Arbor, MI 48109-2122 `hero@eecs.umich.edu`. T. Marzetta is with the Mathematical Sciences Research Center, Bell Laboratories, Lucent Technologies, 600 Mountain Avenue, Murray Hill, NJ 07974 `tlm@research.bell-labs.com`. The work of the first author was partially supported by the Mathematical Sciences Research Center during his sabatical leave there in spring 1999 and subsequently by the Army Research Office under grant ARO DAAH04-96-1-0337. Portions of this work were presented at the IEEE Workshop on Sensors, Arrays and Multichannel Signal Processing (SAM) in Cambridge MA, March 2000, and at the IEEE International Symposium on Information Theory (ISIT) in Sorrento Italy, June 2000.

Figure Captions

1. Two dimensional slice of three dimensional cone cone(E_k) ($k = 3$) with inscribed and circumscribed three dimensional spheres used to establish sufficient and necessary conditions, respectively, for establishing that $\underline{1}_k \in \text{cone}(E_k)$. Interior of triangle is slice of cone(E_k) in a plane orthogonal to the vector $c\underline{u}$. Vertices of triangle are located at points $\{c_i \underline{\varepsilon}_k(i)\}_{i=1}^3$ along extremal rays of cone(E_k). The inner dotted circle is the corresponding slice of the largest possible sphere centered at $c\underline{u}$ that can be inscribed in cone(E_k). The outer dotted circle is the slice of the smallest possible sphere centered at $c\underline{u}$ through which all rays in cone(E_k) must pass.

2. Interior of outer triangle is slice of cone(E_k) for $k = 3$ in a plane orthogonal to the vector $c\underline{1}_k$. Vertices of outer triangle are located at points $\{c_i \underline{\varepsilon}_3(i)\}_{i=1}^3$ along the extremal rays of cone(E_k). The interior of the smaller equilateral triangle in the center is the slice of cone($F_k(\beta_k)$) within the same plane. The equilateral triangle has center point $c\underline{1}_k$. The dotted circle is the slice of a sphere with radius d equal to the distance between $c\underline{1}_k$ and the vertices of the smaller triangle. b is the distance of $c\underline{1}_k$ to the closest face of cone(E_k). cone($F_k(\beta_k)$) $\subset$ cone(E_k) as long as the smaller triangle is inscribed in the larger triangle; i.e. $d \leq b$.

3. Top panel shows M_o given by (55) as a function of the SNR parameter ηTM. Bottom panel is blow up of first panel over a reduced range of SNR. The straight line is a least squares linear fit to the upper panel. The linear approximation has slope 0.32 and zero intercept 0.08. Average residual error between linear fit and exact M_o vs SNR step function is less than 0.09 and maximum error is less than 0.52. By Corollary 2, for T, K, M, M_o satisfying $T \geq KM_o$ and $M_o \leq M$, the curve gives the number of antennas utilized by the optimal constellation for various SNR's

1 Introduction

In this paper we investigate the single-user cut-off rate for a Rayleigh fading spatio-temporal channel model under a maximum peak transmitted power constraint. For a wider perspective on different channel fading models we refer the reader to [4]. Multi-channel flat Rayleigh fading models have been widely used to investigate space-time channel capacity [30, 8, 14, 20, 31, 12], to investigate random coding error exponents [1, 31], and to motivate and evaluate robust space time coding schemes [15, 16, 29, 7]. Similarly to Marzetta and Hochwald [20] our results are derived under the quasi-static Rayleigh fading model and under the assumption that neither transmitter nor receiver knows the channel. The quasi-static fading model assumes a propagation matrix of zero mean complex Gaussian distributed coefficients which remain constant over several time samples, called the coherent fade sampling interval,but change independently over successive fade intervals. The signal set for this channel consists of complex matrices whose rows are indexed over temporal coordinates that span the fade interval, and whose columns are indexed over spatial coordinates that span the locations of the transmitter antennas.

The channel cut-off rate R_o is a lower bound on the Shannon channel capacity C. R_o also specifies an upper bound on the error rate of an optimal decoder operating at a given symbol rate which can be used to majorize the minimum probability of decoding error. Cut-off rate analysis has frequently been adopted to establish practical coding limits [32, 11] as the cut-off rate specifies the highest information rate beyond which sequential decoding becomes impractical [27, 34] and as it is frequently simpler to calculate than channel capacity. Cut-off rate analysis has also been used to evaluate relative merits between different coding and modulation schemes [22], signal design for optical communications [28], and establishing achievable rate regions for multiple access channels [24, 25].

The following are some of the principal results obtained.

1. An integral representation for the cut-off rate is obtained (Lemma 1) which depends on a pairwise dissimilarity measure over the set of signal matrices. This dissimilarity measure is a decreasing function of the spatial correlation between pairs of signal matrices.

2. For low SNR the dissimilarity measure reduces to a distance metric equal to the trace norm of pairwise differences between outerproducts of the signal matrices (Lemma 7).

3. The K dimensional cut-off rate, defined as the cut-off rate for constellations whose dimension does

not exceed K, is the appropriate limiting factor for finite dimensional space time coding/decoding and when signal sets must be generated with finite numerical precision. A necessary condition for such a signal constellation to attain cut-off rate is that there exist a probability distribution on the constellation which equalizes the average exponentiated distance between signal matrices in the constellation (Proposition 3). We call this the equalization condition and it plays a central role in this work.

4. The determination of the K dimensional cut-off rate reduces to maximization of a quadratic form over the set of feasible constellations, defined as those constellations which satisfy both the peak power constraint and the finite dimensional equalization condition (Proposition 4). This quadratic form is similar to that arising in the Capon/MVDR method for adaptive beamforming arrays. If the feasible set of K dimensional constellations is empty then the rate-optimal constellation is necessarily of dimension less than K.

5. For low symbol-rate the rate-optimal constellation is a set of scaled mutually orthogonal unitary matrices in $\mathbb{C}^{T\times M}$ (Proposition 9). This constellation is also distance-optimal in the sense of having the largest possible minimal distance over all constellations of the same dimension (Corollary 2). When SNR is low the rank of the signal matrices in the constellation is one and cut-off rate is achieved by transmitting all power from a single antenna element. As the SNR increases the rank of the signal matrices increases and more and more antenna elements are utilized. Interestingly, the number of receive antennas N plays no role whatsoever in determining how many transmit antennas should be used.

6. When K exceeds T/M but is finite, the K dimensional cut-off rate is not closed form and must be determined by numerical optimization of the cut-off rate objective function over the feasible set of constellations. We derive simple necessary conditions and simple sufficient conditions for a constellation to be in the feasible set (Lemma 3). The necessary conditions characterize properties of the optimal cut-off rate achieving constellation while the sufficient conditions can be used to generate feasible signal sets. To satisfy these conditions, among other attributes, the signal matrices should have pairwise distances of low variability. As these conditions can be used to check if a candidate set of constellations is in the feasible set or not, they provide a potentially useful signal design tool, especially when dealing with large constellations of high dimensional signal matrices for which computing the globally rate-optimal signal set may be impractical.

7. When the number N of receive antennas increases the distance-optimal constellation becomes nearly

rate-optimal and the K dimensional cut-off rate becomes approximately equal to $\ln K$. Furthermore, the dimension K of the rate-optimal constellation among all constellations of countable dimension increases as N increases (Proposition 6).

The outline of this paper is as follows. In Section 2 the Rayleigh fading measurement model is introduced, we briefly review the computational cut-off rate. and we characterize general structural properties of space-time codes that achieve cut-off rate. In Section 3 an integral expression is specified for cut-off rate involving the pairwise dissimilarity function. In Section 4 we specialize to the finite dimensional cut-off rate which reduces to maximization of a quadratic form over signal dissimilarity matrices. In Section 5 we consider quadratic programming for computing the cut-off rate. In Section 6 general characterizations for the feasible set are given. In Section 7 we specialize to the case of finite constellations with dimension smaller than T/M, for which the cut-off rate is closed form.

2 Preliminaries

2.1 Rayleigh Fading Spatio-Temporal Channel

We use substantially the same notation as in the papers of Hochwald and Marzetta [20, 14]. Let there be M transmitter antennas and N receiver antennas and let the MN channel fading coefficients be constant over an interval of length T time periods, called the coherent fade sampling interval. A transmitted signal S is a $T \times M$ matrix having complex valued entries. Let $\mathcal{S} = \mathbb{C}^{T \times M}$ denote the set of all possible signal matrices. The norm of a matrix $S \in \mathcal{S}$ is defined as

$$\|S\| = \sqrt{\mathrm{tr}\{SS^H\}} = \sqrt{\sum_{i=1}^{T} |\lambda_i|^2},$$

where S^H denotes the Hermitian transpose of S and $\{\lambda_i\}_{i=1}^T$ are the singular values of S. Note that if $T > M$ only M of these singular values will be non-zero.

For N receiver antennas and an observation time interval of LT time periods the received signal is a sequence $\{X_i\}_{i=1}^L$ of L complex valued $T \times N$ matrices $X_i \in \mathcal{X} = \mathbb{C}^{T \times N}$ which has the representation [20]:

$$X_i = \sqrt{\eta} S_i H_i + W_i, \quad i = 1, \ldots, L, \tag{1}$$

where $S_i \in \mathcal{S}$ is the i-th transmitted signal, $\eta = \rho/M$ is the normalized signal-to-noise ratio (SNR) with $\rho > 0$ the expected SNR at each receiver per transmit antenna, H_i is an $M \times N$ matrix of complex channel

coefficients, and W_i is a $T \times N$ matrix of complex noises. The quasi-static Rayleigh flat fading model corresponds to taking the $LN(T+M)$ elements of the matrices $\{H_i\}_{i=1}^L$ and $\{W_i\}_{i=1}^L$ to be i.i.d. complex zero mean (circularly symmetric) Gaussian random variables with unit variance. Therefore the joint conditional probability density of the observations factors into a product of marginals

$$p(X_1, \ldots, X_L | S_1, \ldots, S_L) = \prod_{i=1}^{L} p(X_i | S_i),$$

where

$$p(X_i|S_i) = \frac{\exp\left(-\mathrm{tr}\{[I_T + \eta S_i S_i^H]^{-1} X_i X_i^H\right)}{\pi^{TN} \left|I_T + \eta S_i S_i^H\right|^N}, \tag{2}$$

I_T is the $T \times T$ identity matrix, and $|A| = |\det(A)|$ denotes the magnitude determinant of square matrix A.

Let P_e denote the probability of decoding error for a block code of rate R (nats) with blocklength L. It is well known [34, 10, 9] that for R below capacity C, the minimum decoding error probability P_e of the best code satisfies

$$P_e \leq e^{-LE_U(R)}, \tag{3}$$

where $E_U(R) > 0$ is a reliability function, called the random coding error exponent, controlling the error rate of the optimal decoder:

$$E_U(R) = \max_{\mu \in [0,1]} \max_{P \in \mathcal{P}} \left\{ -\mu R - \ln \int_{X \in \mathcal{X}} \left[\int_{S \in \mathcal{S}} \left(p(X|S)\right)^{1/(1+\mu)} dP(S) \right]^{1+\mu} dX \right\}, \quad \text{nats/symbol}, \tag{4}$$

where the inner maximization is performed over a suitably constrained set $\mathcal{P}$ of probability distributions P defined over the set of signal matrices $\mathcal{S}$. Additional constraints on P are determined by factors such as total power budget or envelope constraints and are left implicit in (4). The distribution $P = P^*$ which attains the maximum in $E_U(R)$ gives an optimal signal distribution for which there exists a decoder achieving minimum probability of decoding error for sufficiently large blocklengths. Generally, P^* is not discrete and an optimal set of signal matrices can only be constructed by a random coding procedure. The function $E_U(R)$ was studied for the spatio-temporal Rayleigh quasi-static fading model by Abou-Faycal and Hochwald [1] and Telatar [31] under a mean power constraint on transmitted signal matrices. Unfortunately, the double maximization in (4) is generally very difficult since the inner integral is raised to a fractional exponent when $\mu \in (0, 1)$. The cut-off rate specified below is a lower bound on the error exponent $E_U(R)$ which is frequently simpler to analyze.

The random coding error exponent $E_U(R)$ can be lower bounded by the tangent line $Q(R)$ having slope -1: $Q(R) = R_o - R$, where R_o is this line's intercept on both the $E_U(R)$ and the R axes [34]. The point

of tangency $R_c = \max_{R\in[0,C]}\{R : E_U(R) = Q(R)\}$ is the critical rate and the intercept R_o is called the computational cut-off rate. The cut-off rate is given by

$$R_o = \max_{P\in\mathcal{P}} -\ln \int_{X\in\mathcal{X}} \left[\int_{S\in\mathcal{S}} \sqrt{p(X|S)}dP(S)\right]^2 dX, \quad \text{nats/symbol} \tag{5}$$

which is equivalent to the error exponent $E_U(R)$ in (4) with the maximization over $\mu \in [0,1]$ replaced by the simpler evaluation at the point $\mu = 1$.

While use of cut-off rate for evaluating system performance has its limitations, in particular it is an upper bound only for sequential decoding [5, 21], it satisfies useful properties which in our opinion justify its application to space-time coding:

- Since $R_o \leq C$ cut-off rate provides a lower bound on channel capacity.
- $E_U(R) \approx R_o - R$ when the symbol rate R is close to R_c. Thus the signal distribution $P = P^*$ which attains the maximum in the definition of R_o will attain optimal error rates for R near R_c.
- For sequential decoding strategies, R_o is the maximum practical symbol transmission rate. Specifically, even though Shannon's coding theorems ensure that there exists a low error decoding algorithm for all R such that $R < C$, if $R_o < R < C$ the probability of arbitrary long decision delays becomes significant [27]. Even though it has recently been demonstrated that non-sequential turbo decoders can achieve rates greater than R_o [3], sequential decoding remains of interest and cut-off rate analysis continues to give useful insights [4, 33, 18, 32, 11].
- Combining (3), (4) and (5) a useful upper bound is obtained on the minimum probability of decoder error for transmission rates R less than R_o

  $$P_e \leq \exp\{-L(R_o - R)\}, \quad R < R_o. \tag{6}$$

  Thus $R - R_o$ is an upper bound on the error rate $(\ln P_e)/L$ of the optimal decoder regardless of the decoding algorithm.
- When analysis of channel capacity C is difficult or intractable, R_o offers an alternative which may be easier to analyze.

2.2 Structure of Space-Time Codes Attaining Cut-off Rate

Below we give a result that parallels Theorems 1 and 2 of Marzetta and Hochwald for channel capacity [20], but covers the case of peak power constrained signal sets.

Proposition 1 *Assume that the transmitted signal S is constrained to satisfy the peak power constraint $\|S\|^2 \leq MT$. The peak power constrained cut-off rate attained with $M > T$ transmit antennas is the same as that attained with $M = T$ antennas. Therefore, there is no advantage to using more than T transmit antennas. Furthermore, for $M \leq T$ the signal matrix which achieves peak power constrained cut-off rate can be expressed as $S = V\Lambda$ where V is a $T \times T$ unitary matrix, $\Lambda = [\Lambda_M, O]^T$ is a $T \times M$ matrix, and Λ_M is a diagonal $M \times M$ matrix.*

Proof of Proposition 1

We first treat the case that $M > T$. Let S have singular value decomposition (SVD) $S = V\Lambda U^H$ where V is a $T \times T$ unitary, U is an $M \times M$ unitary and $\Lambda = [\Lambda_T, O]$ is a $T \times M$ matrix partitioned into a diagonal $T \times T$ matrix Λ_T and a $T \times (M - T)$ matrix of zeros. The fading model (1) for $L = 1$ specifies the received signal as the $T \times N$ matrix

$$X = \sqrt{\eta}SH + W,$$

where the $M \times N$ matrix H and the $T \times N$ matrix W are mutually independent matrices of of i.i.d. zero mean complex Gaussian random variables. Substituting the SVD of S into the model we obtain

$$\begin{aligned} X &= \sqrt{\eta}SH + W \\ &= \sqrt{\eta}V\Lambda U^H H + W \\ &= \sqrt{\eta}V\Lambda_T\ [I_T, O]U^H H + W \\ &= \sqrt{\eta}V\Lambda_T \tilde{H}_T + W. \end{aligned}$$

Note that, as U is unitary, $\tilde{H}_T = [I_T, O]U^H H$ is a $T \times N$ matrix of i.i.d. zero mean complex Gaussian r.v.s. Note also that, as $\|S\|^2 = \|V\Lambda_T\|^2$, the $T \times T$ transformed signal $V\Lambda_T$ satisfies the same peak power constraint as the original $T \times M$ signal S. Thus the signal X measured at the receiver after transmission of the signal S on M antennas is statistically equivalent to X received when the signal $V\Lambda_T$ is transmitted on only T antennas.

When $M \leq T$ the SVD of S is identical to $S = V\Lambda U^H$ above except that now $\Lambda = [\Lambda_M, O]^T$, where Λ_M is an $M \times M$ diagonal matrix. Therefore

$$X = \sqrt{\eta}V[\Lambda_M, O]^T \tilde{H} + W$$

where $\tilde{H} = U^H H$. □

Readers familiar with Theorems 1 and 2 of [20] might suspect that characterization of the statistical distribution P of the optimal cut-off achieving signal matrix S can be obtained. Indeed, paralleling the arguments of [20], it can be shown that, as $tr\{SS^H\} \leq TM$ is invariant to unitary premultiplication of S and as the maximization in the definition of R_o is over a concave function of P, the peak-power constrained cut-off rate is attained by random matrices of the form $S = V\Lambda$ where V is a $T \times T$ isotropically distributed matrix, $\Lambda = [\Lambda_M, 0]^T$ is a random $T \times M$ diagonal matrix, and V and Λ are statistically independent.

3 R_o Representation for Quasi-Static Rayleigh Flat Fading Channel

Here we give integral representations for the cut-off rate.

Lemma 1 *The cut-off rate for the spatio-temporal fading model (1) is given by*

$$R_o = \max_{P\in\mathcal{P}} -\ln \int_{S_1\in\mathcal{S}} dP(S_1) \int_{S_2\in\mathcal{S}} dP(S_2) \left[\frac{\sqrt{\left|I_T + \eta S_1 S_1^H\right| \left|I_T + \eta S_2 S_2^H\right|}}{\left|I_T + \frac{\eta}{2}(S_1 S_1^H + S_2 S_2^H)\right|} \right]^N, \quad \text{nats/symbol} \tag{7}$$

where the maximization is performed over the set $\mathcal{P}$ of distributions P of suitably constrained signal matrices.

Proof of Lemma 1

For notational convenience define the $T \times T$ matrix $A(S) = I_T + \eta SS^H$. Then the conditional p.d.f. (2) takes the form $p(X_i|S_i) = \exp\left(-\mathrm{tr}\{A(S_i)^{-1} X_i X_i^H\}\right) / (\pi^{TN} |A(S_i)|^N)$. The integral in (5) is the triple integral $\int dX \int dP(S_1) \int dP(S_2) \sqrt{P(X|S_1)P(X|S_2)}$. As $\eta S_i S_i^H$ is non-negative definite $|A(S_i)| \geq 1$ and therefore $p(X_i|S_i) \leq 1/\pi^{NT} < \infty$. Thus by Fubini we can interchange order of integration in the triple integral to obtain

$$R_o \;=\; \max_{P\in\mathcal{P}} -\ln \int_{S_1\in\mathcal{S}} dP(S_1) \int_{S_2\in\mathcal{S}} dP(S_2) \int_{X\in\mathcal{X}} dX \sqrt{p(X|S_1)p(X|S_2)} \tag{8}$$

The inner integral has the explicit form

$$\begin{aligned} &\int_{X\in\mathcal{X}} dX \sqrt{p(X|S_1)p(X|S_2)} \qquad (9)\\ &\qquad = \frac{1}{|A(S_1)|^{N/2}\,|A(S_2)|^{N/2}} \underbrace{\frac{1}{\pi^{TN}} \int \exp\left(-\tfrac{1}{2}\left[A^{-1}(S_1) + A^{-1}(S_2)\right] XX^H\right)}_{\left|\frac{1}{2}A^{-1}(S_1) + \frac{1}{2}A^{-1}(S_2)\right|^N} \\ &\qquad = \left(\frac{\sqrt{|A(S_1)|\,|A(S_2)|}}{|\frac{1}{2}A(S_1) + \frac{1}{2}A(S_2)|} \right)^N \end{aligned}$$

where in the first line the normalization constant of the multivariate complex Gaussian p.d.f. has been identified ($\int \exp\{-\mathrm{tr}(B^{-1}XX^H)\}dX = \pi^{TN}|B|^N$) and in the second line the determinant property $|BC||B^{-1} + C^{-1}| = |B + C|$ has been used. □

The form of the cut-off rate given in (7) suggests defining the pairwise dissimilarity measure for any two signal matrices $S_1, S_2 \in S$

$$D(S_1\|S_2) \quad \stackrel{\text{def}}{=} \frac{1}{2}\ln\frac{\left|I_T+\frac{\eta}{2}(S_1S_1^H+S_2S_2^H)\right|^2}{\left|I_T+\eta S_1S_1^H\right|\left|I_T+\eta S_2S_2^H\right|}. \tag{10}$$

Thus the cut-off rate can be equivalently expressed as

$$R_o = \max_{P\in\mathcal{P}} -\ln\int_{S_1\in\mathcal{S}} dP(S_1)\int_{S_2\in\mathcal{S}} dP(S_2)\; e^{-ND(S_1\|S_2)} \quad \text{nats/symbol.} \tag{11}$$

As the integral (9) is upper bounded by one (Cauchy-Schwarz inequality) $D(S_1|S_2)$ is non-negative. By Lemma 7, given in the Appendix, $D(S_1\|S_2) = \eta^2/8\|S_1S_1^H - S_2S_2^H\|^2 + o(\eta^2)$, so that in the asymptotic regime of low SNR (η small), $D(S_1\|S_2)$ is to a good approximation proportional to the squared norm $\|S_1S_1^H - S_2S_2^H\|^2$ of the matrix outerproduct difference. We define the minimal distance (dissimilarity) of a set of signal matrices $\mathcal{S}^k = \{S_i\}_{i=1}^k$ as $D_{\min} = \min_{S_i,S_j\in\mathcal{S}^k:i\neq j} D(S_i\|S_j)$. A set of signal matrices which is close to optimal, in terms of nearly attaining the cut-off rate, might be expected to have large value of $D_{\min}$. Indeed, Proposition 9 implies that the rate-optimal peak constrained signal constellation maximizes $D_{\min}$ for low symbol-rates.

In the following sections we specialize to the case of discrete signal constellations for which equalizer distributions are always optimal.

4 Finite (K) Dimensional Cut-off Rate

For practical coding schemes it is of interest to restrict attention to finite sets of signal matrices. Let K be a prespecified finite positive integer. The cut-off rate (11) restricted to discrete distributions concentrated on at most K signal matrices will be called the K dimensional cut-off rate and takes the form

$$\tilde{R}_o(K) \quad = \quad \max_{\{P_i,S_i\}_{i=1}^K\in\mathcal{G}^K} -\ln\tilde{Q}\left(\{P_i\}_{i=1}^K,\{S_i\}_{i=1}^K\right) = -\ln\min_{\{P_i,S_i\}_{i=1}^K\in\mathcal{G}^K}\tilde{Q}\left(\{P_i\}_{i=1}^K,\{S_i\}_{i=1}^K\right) \tag{12}$$

where $\tilde{Q}$ is the quadratic form

$$\tilde{Q}\left(\{P_i\}_{i=1}^K,\{S_i\}_{i=1}^K\right) = \sum_{i=1}^K P_i\sum_{j=1}^K P_j\; e^{-ND(S_i\|S_j)}, \tag{13}$$

and where $\mathcal{G}^K$ is a suitably constrained set of signal matrices $\{S_i\}_{i=1}^K$ and signal probabilities $\{P_i\}_{i=1}^K$ (see subsection below for examples).

When $P_i = 0$ the signal S_i is never transmitted which motivates a natural definition of a signal constellation.

Definition 1 *A set of matrices $\{S_i\}_{i=1}^K$ in $\mathbb{C}^{T\times M}$ is a signal constellation if all assigned signal probabilities P_i are strictly positive, $i = 1, \ldots, K$.*

As the optimization in (12) is performed over a restricted set $\tilde{R}_o(K)$ is a lower bound on R_o. Hence, in particular, $\tilde{R}_o(K)$ can be used in (6) to specify an upper bound on probability of decoding error or a lower bound on capacity. If the optimizing distribution P in (11) over $\mathcal{P}$ actually concentrates on a constellation consisting of a discrete finite number of signal matrices then the bound is tight, i.e. P is a discrete probability and $\tilde{R}_o(K) = R_o$ for some finite K. This hypothesis may not be unreasonable, e.g. when $N = M = 1$, the optimal P was shown to be discrete by Abou-Faycal and Hochwald [1] for codes that achieve the random coding error exponent. However, for cut-off rate and for more general values of N and M this property remains to be verified. On the other hand, when the signal matrices are generated under finite precision arithmetic the bound $\tilde{R}_o(K)$ is always tight for an appropriately chosen K, e.g., $K = 2T \times M2^B$ for B bit register resolution on each complex entry of the signal matrix.

4.1 Peak Power vs. Average Power Constraints

When an average transmitted power constraint is imposed, such as adopted in [20], the optimization (12) must be performed over the restricted set

$$\mathcal{G}^K = \mathcal{G}^K_{\text{avg}} = \left\{ \{P_i, S_i\}_{i=1}^K : P_i \geq 0,\ S_i \in \mathbb{C}^{TM},\ \sum_{i=1}^K P_i = 1,\ \sum_{i=1}^K P_i \|S_i\|^2 \leq TM \right\}$$

where TM is the average transmitter power budget. Thus the average power constraint introduces additional dependency between signal matrices and signal probabilities which complicates the optimization problem.

On the other hand, when a peak power constraint is imposed, the optimization (12) is performed over the simpler product set

$$\mathcal{G}^K = \mathcal{G}^K_{\text{peak}} = \left\{ \{P_i, S_i\}_{i=1}^K : P_i \geq 0, \sum_{i=1}^K P_i = 1, S_i \in \mathbb{C}^{TM}, \|S_i\|^2 \leq TM \right\} = \mathcal{P}^K \times \mathcal{S}^K_{peak} \tag{14}$$

where

$$\mathcal{P}^K = \left\{ \{P_i\}_{i=1}^K : P_i \geq 0, \sum_{i=1}^K P_i = 1 \right\} \tag{15}$$

and

$$\mathcal{S}_{\text{peak}}^K = \left\{ \{S_i\}_{i=1}^K : S_i \in \mathbb{C}^{TM}, \|S_i\|^2 \leq TM \right\}. \tag{16}$$

In particular, under the constraint (14) the optimization (12) decomposes into two nested minimizations

$$\tilde{R}_o(K) = -\ln \min_{\{S_i\}_{i=1}^K \in \mathcal{S}_{\text{peak}}^K} \tilde{Q}\left(\{P_i^*\}_{i=1}^K, \{S_i\}_{i=1}^K\right), \tag{17}$$

where, for fixed $\{S_1\}_{i=1}^K$, $P_i^* = P_i^*(S_1, \ldots, S_K)$, $i = 1, \ldots, K$, is the optimal probability assignment to the signal S_i

$$\{P_i^*\}_{i=1}^K = \text{argmin}_{\{P_i\}_{i=1}^K \in \mathcal{P}^K} \tilde{Q}\left(\{P_i\}_{i=1}^K, \{S_i\}_{i=1}^K\right). \tag{18}$$

4.2 Signal Dissimilarity Matrix

A more compact form for $\tilde{R}_o(K)$ (12) is obtained by putting the quadratic form into vector form with a dissimilarity matrix E_K. For K signal matrices $S_1, \ldots, S_K$, and for $e_K(i,j) = \exp(-ND(S_i \| S_j))$, $E_K = E_K(S_1, \ldots, S_K)$ is the $K \times K$ matrix

$$\begin{aligned} E_K &= ((e_K(i,j)))_{i,j=1}^K \\ &= \begin{bmatrix} 1 & e_{1,2} & \cdots & e_{1,K} \\ e_{2,1} & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & e_{K-1,K} \\ e_{K,1} & \cdots & e_{K,K-1} & 1 \end{bmatrix}. \end{aligned} \tag{19}$$

Note that E_K is symmetric with non-negative entries.

Lemma 2 *E_K is non-negative definite. If the K outerproduct matrices $\{S_i S_i^H\}_{i=1}^K$ are distinct then: (i) E_K is positive definite; and (ii) $e_K(i,j) < 1$, $i \neq j$.*

Proof of Lemma 2

From the defining relations (10) and representation (9) in the proof of Lemma 1 we have $e_K(i,j) = \int_{X \in \mathcal{X}} dX \sqrt{p(X|S_i)p(X|S_j)}$. For assertion (i) consider for any $\underline{a} \in \mathbb{R}^K$

$$\underline{a}^T E_K \underline{a} = \sum_{i=1}^K \sum_{j=1}^K a_i e_K(i,j) a_j$$

$$= \int_{X\in\mathcal{X}} dX \sum_{i=1}^{K}\sum_{j=1}^{K} a_i a_j \sqrt{p(X|S_i)p(X|S_j)}$$
$$= \int_{X\in\mathcal{X}} dX \left(\sum_{i=1}^{K} a_i \sqrt{p(X|S_i)}\right)^2 \geq 0.$$

Thus $\underline{a}^T E_K \underline{a}$ is equal to zero iff $\sum_{i=1}^{K} a_i\sqrt{p(X|S_i)} = 0$ almost everywhere (X). In view of (2): $\sum_{i=1}^{K} a_i\sqrt{p(X|S_i)} = \sum_{i=1}^{K} \tilde{a}_i \tilde{p}(X|S_i)$ where $\tilde{p}(X|S_i)$ is a multivariate circular complex Gaussian density with zero mean and covariance matrix $2[I_T + \eta S_i S_i^H]$ and $\tilde{a}_i = a_i 2^{TN}\pi^{TN/2}|I_T + \eta S_i S_i^H|^{N/2}$. For a reference signal $S_1 \in \mathcal{S}$ an easy calculation establishes that for any $S_2 \in \mathcal{S}$ the Kullback-Leibler divergence [6] of $\tilde{p}(X|S_2)$ from $\tilde{p}(X|S_1)$ is

$$\begin{aligned} K(S_1\|S_2) &= \int_{X\in\mathcal{X}} \tilde{p}(X|S_1)\ln\left(\tilde{p}(X|S_1)/\tilde{p}(X|S_2)\right) dX \\ &= N\sum_{i=1}^{K}[(\sigma_i - 1) - \ln\sigma_i] \end{aligned}$$

where $\{\sigma_i\}_{i=1}^{K}$ are the eigenvalues of the positive definite matrix $A = [I_T + \eta S_2 S_2^H]^{-\frac{1}{2}}[I_T + \eta S_1 S_1^H][I_T + \eta S_2 S_2^H]^{-\frac{1}{2}}$. As $(\sigma_i - 1) - \ln\sigma_i \geq 0$, with equality iff $\sigma_i = 1$, we see that $K(S_1\|S_2) = 0$ iff $A = I_K$, as A is symmetric. Thus $K(S_1\|S_2) = 0$ iff $S_1 S_1^H = S_2 S_2^H$ and therefore for distinct $\{S_i S_i^H\}_{i=1}^{K}$ the density functions $\{\tilde{p}(X|S_i)\}_{i=1}^{K}$ are almost everywhere linearly independent functions of X [17]. Gathering together the above results: if $\{S_i S_i^H\}_{i=1}^{K}$ are distinct then $\sum_{i=1}^{K} a_i\sqrt{p(X|S_i)} = 0$ (a.e.) implies $\{a_i\}_{i=1}^{K}$ are identically zero. Hence E_K is positive definite.

As for assertion (ii) of the Lemma we have by Cauchy-Schwarz

$$\begin{aligned} e_K^2(i,j) &= \left(\int_{X\in\mathcal{X}} dX\sqrt{p(X|S_i)p(X|S_j)}\right)^2 \\ &\leq \int_{X\in\mathcal{X}} p(X|S_i)dX \int_{X\in\mathcal{X}} p(X|S_j)dX \end{aligned}$$

with equality iff $p(X|S_i) = p(X|S_j)$ (a.e. X). As shown above, this equality condition can only occur if $S_i S_i^H = S_j S_j^H$. Hence $e_K(i,j) < 1$ under the hypotheses of the lemma. □

5 General Properties of Solutions to R_o

In this and the following subsections we establish properties of solutions to (12) under the peak power constraint (14).

First we give a relation between the average distance of the rate-optimal constellation and the largest

possible minimum distance of any constellation of identical dimension. Define this latter distance

$$D^{**}_{\min} \stackrel{\text{def}}{=} \max_{\{S_i\}_{i=1}^K \in \mathcal{S}^K_{\text{peak}}} \min_{i \neq j} D(S_i \| S_j).$$

A constellation whose minimum distance $D_{\min}$ attains $D^{**}_{\min}$ is said to be distance-optimal.

Proposition 2 *Let $\{S_i^*\}_{i=1}^K$ be a constellation that attains the K dimensional cut-off rate $\tilde{R}_o(K)$. Then,*

$$\overline{D(S_i^* \| S_j^*)} \geq D^{**}_{\min},$$

where $\overline{D(S_i^* \| S_j^*)}$ is the mean value

$$\overline{D(S_i^* \| S_j^*)} = \frac{\sum_{i \neq j} P_i^* P_j^* D(S_i^* \| S_j^*)}{\sum_{i \neq j} P_i^* P_j^*}.$$

As $\overline{D(S_i^* \| S_j^*)} \leq \max_{i \neq j} D(S_i^* \| S_j^*)$, combination of Proposition 2 and a sphere-packing bound on $D^{**}_{\min}$ derived in [13, Prop 2] gives the following bound on the maximum distance of the constellation attaining cut-off rate

$$\max_{i \neq j} D(S_i^* \| S_j^*) \geq \frac{\eta^2 (TM)^2}{128} \, K^{-2/T} \, (1 + o(\eta^2)).$$

Proof of Proposition 2

Let $\{P_i^*\}_{i=1}^K$ achieve the minimum value $\tilde{Q}^*$ of the quadratic form $\tilde{Q}\left(\{P_i\}_{i=1}^K, \{S_i^*\}_{i=1}^K\right)$ in (18). First we show that

$$\frac{\sum_{i \neq j} P_i^* P_j^* e^{-N D(S_i^* \| S_j^*)}}{\sum_{i \neq j} P_i^* P_j^*} \leq e^{-N D^{**}_{\min}}, \tag{20}$$

Assume the contrary. Then

$$\begin{aligned} \tilde{Q}^* &= \sum_{i \neq j}^K P_i^* P_j^* e^{-N D(S_i^* \| S_j^*)} + \sum_{i=1}^K (P_i^*)^2 \\ &> e^{-N D^{**}} \sum_{i \neq j}^K P_i^* P_j^* + \sum_{i=1}^K (P_i^*)^2. \end{aligned}$$

On the other hand,

$$\begin{aligned} \tilde{Q}^* &\leq \min_{\{S_i\}_{i=1}^K \in \mathcal{S}^K_{\text{peak}}} \left\{ \sum_{i \neq j}^K P_i^* P_j^* e^{-N D(S_i \| S_j)} + \sum_{i=1}^K (P_i^*)^2 \right\} \\ &\leq \min_{\{S_i\}_{i=1}^K \in \mathcal{S}^K_{\text{peak}}} \left\{ e^{-N \min_{i \neq j} D(S_i \| S_j)} \right\} \sum_{i \neq j}^K P_i^* P_j^* + \sum_{i=1}^K (P_i^*)^2 \\ &= e^{-N D^{**}_{\min}} \sum_{i \neq j}^K P_i^* P_j^* + \sum_{i=1}^K (P_i^*)^2, \end{aligned}$$

which is a contradiction. The Proposition follows after application of Jensen's inequality to the left hand side of (20).

5.1 Optimality of Equalizer Distributions

The solution $\underline{P}_K^* = [P_1^*, \ldots, P_K^*]^T$ to the minimization (18) can be found by convex quadratic optimization subject to linear equality and inequality constraints. Define the Lagrangian

$$J(\underline{P}_K) \quad = \quad \underline{P}_K^T E_K \underline{P}_K - 2c(\underline{1}_K^T \underline{P}_K - 1), \tag{21}$$

where, $\underline{P}_K = [P_1, \ldots, P_K]^T$, $\underline{1}_K = [1, \ldots, 1]^T$ is a K-element vector of ones, and $c > 0$ is an undetermined multiplier that must be chosen to enforce the equality constraint $\underline{1}_K^T \underline{P} = 1$. As $J(\underline{P})$ is convex cup the Kuhn-Tucker conditions [26] assert that the minimum exists and must satisfy

$$\frac{\partial J(\underline{P}_K^*)}{\partial P_i^*} \quad \begin{cases} = 0 & \text{if} \quad P_i^* > 0 \\ \geq 0 & \text{if} \quad P_i^* = 0 \end{cases}$$

or equivalently

$$\underline{\varepsilon}_K^T(i) \underline{P}_K^* \quad = \quad c \quad \text{if} \quad P_i^* > 0 \tag{22}$$

$$\underline{\varepsilon}_K^T(i) \underline{P}_K^* \quad \geq \quad c \quad \text{if} \quad P_i^* = 0 \tag{23}$$

where $\underline{\varepsilon}_K(i)$ is the i-th column of E_K. By reordering the entries of $\underline{P}_K^*$ and the rows and columns of E_K we can assume, without loss of generality, that the $\underline{\varepsilon}_K(i)$,'s satisfying (22) are the first k columns of E_K with the remainder satisfying (23), $k \leq K$. Making this assumption, (22) and (23) imply that $P_{k+1}^* = \ldots = P_K^* = 0$ and

$$E_k \underline{P}_k^* = c \underline{1}_k. \tag{24}$$

Furthermore, we can assume that $\{S_i S_i^H\}_{i=1}^K$ are distinct (see Lemma 9) and therefore, by Lemma 2, E_k is positive definite. Therefore, from (22)-(24) the strictly positive component of the minimizer $\underline{P}_K^*$ is given by

$$\underline{P}_k^* = c E_k^{-1} \underline{1}_k, \tag{25}$$

where c is determined via the constraint $\underline{1}_K^T \underline{P}_K^* = \underline{1}_k^T \underline{P}_k^* = 1$, or

$$c = 1 / \underline{1}_k^T E_k^{-1} \underline{1}_k. \tag{26}$$

Note that $c > 0$ since E_k is positive definite and $c \leq 1$ since $(\underline{1}_k^T E_k^{-1} \underline{1}_k)(\underline{1}_k^T E_k \underline{1}_k) \geq (\underline{1}^T \underline{1})^2 = k^2$ and $\underline{1}^T E_k \underline{1} \leq k^2$ as the elements of E_k are ≤ 1. Thus, in view of (24), the strictly positive elements of $\underline{P}_K^*$ satisfy

the equalization condition

$$\sum_{j=1}^{k} P_j^* e^{-ND(S_j\|S_i)} = c, \qquad i = 1, \ldots, k, \tag{27}$$

where $\{S_i\}_{i=1}^k$ are the signal matrices in $\{S_i\}_{i=1}^K$ with strictly positive assigned probabilities $\{P_i^*\}_{i=1}^k$. For low SNR η it is easily shown using Lemma 7 that the equalization condition (27) is equivalent to equalizing over $i = 1, \ldots, k$ the average distance from S_i to all other codewords S_j, $i \neq j$:

$$\sum_{j \neq i} P_j^* D(S_j\|S_i) = \alpha + o(\eta^2), \qquad i = 1, \ldots, k,$$

where $\alpha = (1-c)/N$.

We have thus shown that rate-optimal finite dimensional constellations, defined in Definition 1, must have equalizer distributions. These results are summarized in the following.

Proposition 3 *A constellation of dimension K achieves the K dimensional peak constrained cut-off rate (12) only if the optimal distribution $\{P_i^*\}_{i=1}^K$ over signal matrices in the constellation is an equalizer distribution of the form $\underline{P}_K^* = E_K^{-1}\underline{1}_K / \underline{1}_K^T E_K^{-1} \underline{1}_K$.*

An equivalent condition to (24) is that there exist a vector $\underline{x} = [x_1, \ldots, x_k]^T$, not identically zero, lying in the positive orthant $\mathbf{R}_+^k$ which satisfies

$$E_k \underline{x} = \underline{1}_k, \tag{28}$$

or, equivalently, $E_k^{-1}\underline{1}_k \in \mathbf{R}_+^k$. Define the feasibility set $\tilde{\mathcal{S}}_{\text{peak}}^K$ of K-dimensional constellations

$$\tilde{\mathcal{S}}_{\text{peak}}^K = \left\{ \{S_i\}_{i=1}^K : S_i \in \mathbb{C}^{TM},\ \|S_i\|^2 \leq TM, E_K^{-1}\underline{1}_K \in \mathbf{R}_+^K,\ S_i S_i^H \neq S_j S_j^H, i \neq j \right\}. \tag{29}$$

By Proposition 3 $\tilde{R}_o(K)$ is attained by a constellation of dimension K only if $\tilde{\mathcal{S}}_{\text{peak}}^K$ is non-empty. Substitution of the form of the optimal probability vector $\underline{P}_K^*$ specified in Proposition (3) into (13) we obtain the following alternative characterization of $\tilde{R}_o(K)$.

Proposition 4 *Let K be a positive integer. The peak power constrained K dimensional cut-off rate is*

$$\tilde{R}_o(K) \quad = \quad \ln \left\{ \max_{0<k\leq K} \max_{\{S_i\}_{i=1}^k \in \tilde{\mathcal{S}}_{\text{peak}}^k} \underline{1}_k^T E_k^{-1} \underline{1}_k \right\}. \tag{30}$$

Observe that by taking the limit $K \to \infty$ in Proposition 4, we obtain the cut-off rate of constellations of countable, but possibly infinite, dimension. The objective function $\underline{1}_k^T E_k^{-1} \underline{1}_k$ maximized in (30) is similar to the criterion used in Capon's method, also known as minimum variance distortionless response (MVDR), for adapting the weights of a beamforming array of antenna elements and for high resolution spectral estimation [19].

The cut-off rate $\tilde{R}_o(K)$ and the rate achieving signal set can be iteratively computed using (30) in the following steps: (1) select a candidate set of signal matrices $\{S_i\}_{i=1}^K$ in $\mathcal{S}_{\text{peak}}^K$, (2) use quadratic programming to find the zero entries of $\underline{P}_K^*$, specified by condition (23), e.g. using slack variable or active set methods [23]; (3) reorder the zero and non-zero entries of $\underline{P}_K^*$ and permute the rows and columns of E_K as described above; (4) solve $E_K \underline{x} = \underline{1}_K$ for $\underline{x}$; (5) compute inner product $\underline{1}^T \underline{x}$ and take its natural logarithm; (6) perturb the candidate set of signal matrices and repeat (1)-(5). This iterative procedure is repeated for each new set of candidate signal matrices until $1/\underline{1}_K^T E_K^{-1} \underline{1}_K$ attains a maximum. The procedure becomes computationally heavy when K becomes large. Proposition 4 suggests an alternative optimization method which bypasses steps (1) and (2) by preselecting the candidate set of signal matrices $\{S_i\}_{i=1}^K$ to lie in $\tilde{\mathcal{S}}_{\text{peak}}^K$.

It will thus be of interest to establish conditions on E_K which guarantee existence of a positive vector $\underline{x}$ satisfying (28) or, equivalently, which guarantee that the vector $E_K^{-1} \underline{1}_K$ lies in the positive orthant. The first result in this direction is that this always holds for a sufficiently large number N of receiver antennas.

5.2 Large Number of Receiver Antennas

The equalization condition (27) is easily manipulated to yield $P_i^* + O(e^{-N D_{\min}}) = c$, $i = 1, \ldots, k$. This suggests that, for any constellation having $D_{\min} > 0$, if N is sufficiently large the equiprobable distribution $P_i = 1/K$, $i = 1, \ldots, K$ will satisfy this condition. We give stronger results below.

Proposition 5 *Let $\{S_i\}_{i=1}^K$ be a set of peak constrained signal matrices with distinct outerproducts $\{S_i S_i^H\}_{i=1}^K$. Define the finite positive integer N_o*

$$N_o = \lfloor \frac{\ln(K-1)}{D_{\min}} \rfloor + 1, \tag{31}$$

where $D_{\min} = \min_{i \neq j} D(S_i \| S_j)$. Then, for $N \geq N_o$ the optimal distribution defined in (18) is

$$\underline{P}_K^* = E_K^{-1} \underline{1}_K / \underline{1}_K^T E_K^{-1} \underline{1}_K = (1 + O(\delta^{N - N_o + 1}))\, \underline{1}_K / K, \tag{32}$$

where $\delta = \exp(-D_{\min})$.

The significance of the above proposition is that if K is no greater than $\exp(ND_{\min}) + 1$ any peak constrained set of signal matrices $\{S_i\}_{i=1}^K$ is in the feasible set $\tilde{\mathcal{S}}_{\text{peak}}^K$ and thus has an equalizer distribution $\underline{P}^*(S_1, \ldots, S_K)$. Furthermore, as $N \to \infty$, $\underline{P}_K^*$ converges to the equiprobable distribution $\underline{1}_K/K$. Using [13, Prop. 2] the following bound on N_o for the distance-optimal constellation can be derived: $N_o \leq \lfloor 128K^{2/T} \ln(K-1)/(\eta^2(TM)^2) \rfloor + 1$ (η^2 small).

Proof of Proposition 5

First observe that under the hypothesis of the proposition $N_o < \infty$ since, by Lemma 2, $D(S_i \| S_j) > 0$, $i \neq j$. Define the symmetric matrix $\Delta = E_K - I_K$. The elements $\{d_{i,j}\}$ of Δ are $d_{i,j} = e^{-ND(S_i \| S_j)} \leq \delta^N$, $i \neq j$, and $d_{i,i} = 0$. Therefore $0 \leq d_{i,j} < 1$. When $N \geq N_o$ the eigenvalues σ_i^Δ of Δ are bounded

$$|\sigma_i^\Delta| \leq \delta^N(K-1) < 1, \quad i = 1, \ldots, K. \tag{33}$$

The left hand inequality of (33) follows from the sequence of inequalities

$$\begin{aligned}
\max_i |\sigma_i^\Delta| &\leq \max_{\underline{y} \in \mathbb{R}^K} \frac{|\underline{y}^T \Delta \underline{y}|}{\underline{y}^T \underline{y}} = \max_{\underline{y} \in \mathbb{R}^K} \frac{\left| \sum_{i,j=1}^K y_i y_j d_{i,j} \right|}{\sum_{i=1}^K y_i^2} \\
&\leq \max_{\underline{y} \in \mathbb{R}^K} \frac{\left| \sum_{i=1}^K y_i^2 \sum_{j=1}^K d_{i,j} \right|}{\sum_{i=1}^K y_i^2} \\
&\leq \max_i \sum_{j=1}^K d_{i,j} \\
&\leq \delta^N(K-1).
\end{aligned}$$

The inequality on the second line follows from Cauchy-Schwarz

$$\left(\sum_{i,j=1}^K (y_i \sqrt{d_{i,j}})(y_j \sqrt{d_{i,j}}) \right)^2 \leq \left(\sum_{i,j=1}^K (y_i \sqrt{d_{i,j}})^2 \right)^2,$$

and the inequality on the third line follows from non-negativity of $d_{i,j}$. The right hand inequality of (33) follows from

$$\delta^N(K-1) = \delta^{N_o}(K-1)\delta^{N-N_o} \leq \delta^{N-N_o+1} < 1, \quad N \geq N_o. \tag{34}$$

Now (28) can be expressed as the perturbed linear system by expressing $E_K = I_K + \Delta$

$$[I_K + \Delta]\underline{x} = \underline{1}_K.$$

As E_K is positive definite $[I_K + \Delta]$ is invertible. Furthermore, from (33) the eigenvalues of Δ^2 are strictly less than one and by elementary matrix manipulations

$$\underline{x} = [I_K + \Delta]^{-1}\underline{1}_K = [I_K - \Delta^2]^{-1}[I_K - \Delta]\underline{1}_K.$$

The i-th element of the vector $[I_K - \Delta]\underline{1}_K$ is $2 - \sum_{j=1}^K d_{i,j} \geq 1 - (K-1)\delta^N$ which by (34) is positive for $N \geq N_o$. Also, $[I_K - \Delta^2]^{-1} = I_K + \sum_{m=1}^\infty \Delta^{2m}$ has positive entries. This implies that for $N \geq N_o$, $\underline{x} \in \mathbb{R}_+^K$ and the i-th element of $\Delta\underline{1}_K$ is upper bounded by $(K-1)\delta^N$ and consequently, as it is non-negative, of order $O(\delta^{N-N_o+1})$ by (34). Hence, $\underline{x} = [I_K + \Delta]^{-1}\underline{1}_K = \underline{1}_K + O(\Delta\underline{1}_K) = \underline{1}_K(1 + O(\delta^{N-N_o+1}))$. □

From the proof of Proposition 5 we can identify a weaker condition on the number of receiver antennas necessary to ensure $\underline{P}_K^*$ be an equalizer distribution: N be sufficiently large to make E_K a diagonally dominant matrix, i.e. to make $\sum_{j=1, j\neq i}^K e^{-ND(S_i\|S_j)} < 1$, $i = 1, \ldots, K$.

For large K a significantly stronger version of Proposition 5 can be established. For arbitrary signal matrices $\{S_i\}_{i=1}^K$ and probabilities $\{P_i\}_{i=1}^K$ define the quantity

$$\tilde{R}_o(K, \{S_i, P_i\}_{i=1}^K) = -\ln \sum_{i,j=1}^K P_i P_j e^{-ND(S_i\|S_j)}.$$

With this definition and (12) we observe that the cut-off rate is expressed as $\tilde{R}_o(K) = \tilde{R}_o(K, \{S_i^*, P_i^*\}_{i=1}^K)$ where, as above, $\{S_i^*, P_i^*\}_{i=1}^K$ is the rate-optimal set of signals and probabilities.

Proposition 6 *Fix K and assume $N \geq \lfloor \frac{\ln(K-1)}{D_{\min}^{**}} \rfloor + 1$. Then*

$$\ln(K) - \ln 2 \leq \tilde{R}_o(K, \{S_i, P_i\}_{i=1}^K) \leq \tilde{R}_o(K) \leq \ln K \tag{35}$$

where $\{S_i\}_{i=1}^K$ is the distance-optimal constellation defined earlier in Section 5, $D_{\min}^{**} = D_{\min}^{**}(K) = \min_{i\neq j} D(S_i\|S_j)$ is its minimal distance, and $P_i = 1/K$ corresponds to the equiprobable distribution. Furthermore, as N increases to infinity the size of any rate-optimal countable constellation and the cut-off rate both increase without bound.

The above Proposition implies that for a sufficiently large number N of receive antennas the K-dimensional cut-off rate takes the form $\tilde{R}_o(K) = \ln K - O(1)$ which is attained by an equiprobable K-dimensional distance-optimal constellation.

Proof of Proposition 6

The upper bound $\tilde{R}_o(K) \leq \ln K$ holds independently of N and is an immediate consequence of the inequality

$$\left(\sum_{i=1}^K \sqrt{p(X|S_i)} P_i\right)^2 \geq \sum_{i=1}^K p(X|S_i) P_i^2$$

which when substituted into the discrete form of (5) yields

$$\begin{aligned}\tilde{R}_o(K) &= \max_{\{P_i\}_{i=1}^K} \max_{\{S_i\}\in\mathcal{S}_{\text{peak}}^K} -\ln \int_{X\in\mathcal{X}} \left[\sum_{i=1}^K \sqrt{p(X|S_i)}P_i\right]^2 dX \\ &\leq \max_{\{P_i\}_{i=1}^K} -\ln \sum_{i=1}^K P_i^2 \\ &\leq \ln K\end{aligned}$$

where the last line follows from the elementary inequality $\sum_{i=1}^K P_i^2 \geq 1/K$, for any set of probabilities $\{P_i\}_{i=1}^K$, with equality when $P_i = 1/K$.

As for the two lower bounds on $\tilde{R}_o(K)$ in (35) first observe that, with S_i and P_i as defined in Proposition 6,

$$\begin{aligned}\tilde{R}_o(K) &\geq \tilde{R}_o(K, \{S_i, P_i\}_{i=1}^K) \\ &= -\ln \frac{1}{K^2} \sum_{i,j=1}^K e^{-ND(S_i\|S_j)} \\ &\geq -\ln \frac{1}{K}\left(1 + (K-1)e^{-ND_{\min}^{**}(K)}\right),\end{aligned}$$

as $D(S_i\|S_i) = 0$ and $D(S_i\|S_j) \geq D_{\min}^{**}(K)$, $i \neq j$. Therefore, since

$$N \geq \lfloor \frac{\ln(K-1)}{D_{\min}^{**}(K)} \rfloor + 1 \geq \frac{\ln(K-1)}{D_{\min}^{**}(K)}$$

we have

$$-\ln \frac{1}{K}\left(1 + (K-1)e^{-ND_{\min}^{**}(K)}\right) \geq -\ln \frac{2}{K} = \ln K - \ln 2$$

and the two lower bounds in (35) are established. Furthermore, for arbitrary $N > 0$ let

$$K \stackrel{\text{def}}{=} \max\left\{k : \lfloor \frac{\ln(k-1)}{D_{\min}^{**}(k)} \rfloor + 1 \leq N\right\}.$$

Then, as $D_{\min}^{**}(k+1) \leq D_{\min}^{**}(k)$ both K and $\tilde{R}_o(K) = \ln(K) - O(1)$ are monotone increasing in N as $N \to \infty$. □

6 Characterization of Feasible Set of Constellations

Here we derive sufficient conditions and necessary conditions for $E_K^{-1}\underline{1}_K$ to lie in the positive orthant. Several of these conditions will be defined recursively in K. So as to not confuse the reader, and to improve clarity of the equations, we will use lower case k throughout this section to distinguish it from the fixed index K.

For a square $k \times k$ matrix A consisting of columns $\underline{a}_1, \ldots, \underline{a}_k$ the positive polyhedral cone generated by A is [2, 26]

$$\text{cone}(A) = \{A\underline{x} : \underline{x} \in \mathbb{R}_+^k\} = \left\{ \sum_{i=1}^{k} \underline{a}_i x_i : x_i > 0,\ i = 1, \ldots, k \right\}.$$

This cone is convex and each of its k faces are contained in one of the hyper-planes $\text{span}\{\underline{a}_1, \ldots, \underline{a}_{i-1}, \underline{a}_{i+1}, \ldots, \underline{a}_k\}$. The extremals of $\text{cone}(A)$ are the k positive rays $\{c\underline{a}_i : c > 0\}$, $i = 1, \ldots, k$.

It is evident that if $\text{cone}(E_k)$ contains the vector $\underline{1}_k$ then, as E_k is positive definite, $\underline{x} = E_k^{-1}\underline{1}$ must have positive elements, which would guarantee that $\underline{P}_k^* = \underline{x}/\underline{1}_k^T\underline{x}$ is the optimal equalizer distribution.

6.1 Conditions for Feasibility

Here we give two conditions, one sufficient and one necessary, to ensure $\underline{1}_k \in \text{cone}(E_k)$. The sufficient condition reduces to specifying the largest inscribed right circular cone which fits inside $\text{cone}(E_k)$ while the necessary condition is equivalent to specifying the smallest right circular cone which contains $\text{cone}(E_k)$. This sufficient condition is slightly weaker than the sufficient condition presented in Lemma 4 of the next section and obtained by recursive inscription of a polyhedral cone. On the other hand, as contrasted with the condition of Lemma 4, the sufficient condition in this section is not easy to express recursively in k

As in (22)-(23), $\underline{\varepsilon}_k(i)$ will denote the i-th column of $E_k = ((e_k(i,j)))_{i,j=1}^k$. For $i = 1, \ldots, k$, define $E_k(-i)$ the $k \times (k-1)$ matrix obtained by deleting its i-th column. Define

$$\Pi_{E_k(-i)} = E_k(-i)[E_k^T(-i)E_k(-i)]^{-1}E_k^T(-i) \tag{36}$$

the idempotent $k \times k$ matrix which orthogonally projects vectors in $\mathbb{R}^k$ onto the column span of $E_k(-i)$.

Lemma 3 *Let $\underline{u}$ be any vector in $\text{cone}(E_k)$. A sufficient condition for $\underline{1}_k$ to be contained in $\text{cone}(E_k)$ is*

$$\left\| \underline{u} - \frac{\underline{u}^T\underline{1}_k}{\|\underline{1}_k\|^2}\underline{1}_k \right\|^2 < \min_i \underline{u}^T[I - \Pi_{E_k(-i)}]\underline{u}. \tag{37}$$

A necessary condition for $\underline{1}_k$ to be contained in $\text{cone}(E_k)$ is

$$\left\| \underline{u} - \frac{\underline{u}^T\underline{1}_k}{\|\underline{1}_k\|^2}\underline{1}_k \right\|^2 < \max_i \left\| \underline{u} - \frac{\underline{u}^T\underline{\varepsilon}_k(i)}{\|\underline{\varepsilon}_k(i)\|^2}\,\underline{\varepsilon}_k(i) \right\|^2. \tag{38}$$

Both of the conditions specified in Lemma 3 require that the test vector $\underline{u}$ be close to the ray $\{c\underline{1}_k : c > 0\}$. As a particularly simple application of the Lemma, consider setting $\underline{u} = \overline{\underline{\varepsilon}}_k = E_k\underline{1}_k/k$; the arithmetic mean

of the columns of E_k. In this case, as all but the i-th column of E_k are orthogonal to $I_k - \Pi_{E_k(-i)}$, the sufficient condition becomes

$$\left\| \overline{\underline{\varepsilon}}_k - \frac{\overline{\underline{\varepsilon}}_k \underline{1}_k}{\|\underline{1}_k\|^2} \underline{1}_k \right\|^2 < \frac{1}{k^2} \min_i \underline{\varepsilon}_k^T(i)[I - \Pi_{E_k(-i)}]\underline{\varepsilon}_k(i).$$

This condition requires that the distance between $\underline{1}_k$ and $\overline{\underline{\varepsilon}}_k$ be less than k^{-2} times the prediction error squared of the most predictable column of E_k using the remaining columns of E_k as predictor vectors.

Proof of Lemma 3

Fix $c > 0$ and let $\underline{z} = c\underline{u}$. As $\underline{z} \in \text{cone}(E_k)$ there exists a sphere centered at $\underline{z}$ that fits inside $\text{cone}(E_k)$ (See Figure 1). Such a sphere can have radius b no larger than the distance between $\underline{z}$ and the closest face of $\text{cone}(E_k)$. As the k faces of $\text{cone}(E_k)$ are contained in the column span of $E_k(-i)$, $i = 1, \ldots, k$ and $\text{cone}(E_k)$ is convex, the radius b of the largest sphere is given by the distance between $\underline{z}$ and its orthogonal projection onto this span

$$\begin{aligned} b^2 &= \min_i \left\{ \min_{\underline{\hat{y}} \in \text{span}\{E_k(-i)\}} \left\{ \|\underline{z} - \underline{\hat{y}}\|^2 \right\} \right\} \\ &= \min_i \underline{z}^T [I - \Pi_{E_k(-i)}]\underline{z} \\ &= c^2 \min_i \underline{u}^T [I - \Pi_{E_k(-i)}]\underline{u}, \end{aligned}$$

which is the right hand side of (37). If the positive ray $\{a\underline{1}_k : a > 0\}$ passes through this sphere then, by convexity of $\text{cone}(E_k)$, $\underline{1}_k$ must lie in $\text{cone}(E_k)$. This occurs iff the distance from $\underline{z}$ to this ray is less than b^2, i.e.

$$\left\| \underline{z} - \frac{\underline{z}^T \underline{1}_k}{\|\underline{1}_k\|^2} \underline{1}_k \right\|^2 = c^2 \left\| \underline{u} - \frac{\underline{u}^T \underline{1}_k}{\|\underline{1}_k\|^2} \underline{1}_k \right\|^2 < b^2$$

Since $c > 0$ is arbitrary, we obtain the inequality (37).

As for the necessary condition, for any $c > 0$ there exists a sphere centered at $\underline{z} = c\underline{u}$ through which all positive rays in $\text{cone}(E_k)$ must pass (See Figure 1). The smallest such sphere has radius equal to the maximum distance between $\underline{z}$ and the faces of $\text{cone}(E_k)$. Points at maximum distance must occur along one of the rays $\{a\underline{\varepsilon}_k(i) : a > 0\}$, $i = 1, \ldots, k$, which are the extremals of $\text{cone}(E_k)$. Therefore the radius of this smallest sphere is

$$d^2 = \max_i \left\| \underline{z} - \frac{\underline{z}^T \underline{\varepsilon}_k(i)}{\|\underline{\varepsilon}_k(i)\|^2} \underline{\varepsilon}_k(i) \right\|^2 = c^2 \max_i \left\| \underline{u} - \frac{\underline{u}^T \underline{\varepsilon}_k(i)}{\|\underline{\varepsilon}_k(i)\|^2} \underline{\varepsilon}_k(i) \right\|^2.$$

Finally, if $\underline{1}_k \in \text{cone}(E_k)$ this sphere must intersect the ray $\{a\underline{1}_k : a > 0\}$ in which case $c^2 \left\| \underline{u} - (\underline{u}^T \underline{1}_k)/\|\underline{1}_k\|^2 \, \underline{1}_k \right\|^2 < d^2$. □

6.2 A Recursive Construction

Here the objective will be to specify conditions for which $\underline{1}_{k+1}$ lies in cone(E_{k+1}) when it is known that $\underline{1}_k$ lies in cone(E_k). To proceed we will need a recursive update for $E_{k+1}^{-1}\underline{1}_{k+1}$ in terms of $E_k^{-1}\underline{1}_k$. For a set of signal matrices $S_1, \ldots, S_{k+1}$ let the $(k+1) \times (k+1)$ matrix E_{k+1}be partitioned as

$$E_{k+1} = \begin{bmatrix} E_k & \underline{e}_k \\ \underline{e}_k^T & 1 \end{bmatrix}, \tag{39}$$

where $\underline{e}_k$ is the vector of pairwise dissimilarity functions

$$\underline{e}_k = \left[e^{-ND(S_1\|S_{k+1})}, \ldots, e^{-ND(S_k\|S_{k+1})}\right]^T. \tag{40}$$

We also recall the partitioned matrix inverse identity for the case that E_{k+1} is positive definite

$$E_{k+1}^{-1} = \begin{bmatrix} E_k^{-1} + \frac{1}{|E_{k+1}|}E_k^{-1}\underline{e}_k\underline{e}_k^T E_k^{-1} & -\frac{1}{|E_{k+1}|}E_k^{-1}\underline{e}_k \\ -\frac{1}{|E_{k+1}|}\underline{e}_k^T E_k^{-1} & \frac{1}{|E_{k+1}|} \end{bmatrix} \tag{41}$$

where $|E_{k+1}| = 1 - \underline{e}^T E_k^{-1}\underline{e} > 0$. Let $\underline{x}^k = E_k^{-1}\underline{1}_k$ have partitioned form $\underline{x}^k = [\underline{x}_{k-1}^k, x_k^k]^T$ where x_k^k is a scalar.

Proposition 7 *For a given set of signal matrices $\{S_i\}_{i=1}^{k+1}$ with distinct outerproducts $\{S_iS_i^H\}_{i=1}^{k+1}$*

$$\begin{bmatrix} \underline{x}_k^{k+1} \\ x_{k+1}^{k+1} \end{bmatrix} = \begin{bmatrix} E_k^{-1}(\underline{1}_k - \alpha_k\underline{e}_k) \\ \alpha_k \end{bmatrix} = \begin{bmatrix} \underline{x}^k - \alpha_k E_k^{-1}\underline{e}_k \\ \alpha_k \end{bmatrix}, \tag{42}$$

where

$$\alpha_k = \frac{1 - \underline{e}_k^T E_k^{-1}\underline{1}_k}{1 - \underline{e}_k^T E_k^{-1}\underline{e}_k}. \tag{43}$$

Proof of Proposition 7

By Lemma 2 E_{k+1} is positive definite. Applying the partitioned matrix inverse identity (41) to the right hand side of $\underline{x}^{k+1} = E_{k+1}^{-1}\underline{1}_{k+1}$,

$$\begin{aligned} E_{k+1}^{-1}\underline{1}_{k+1} &= \begin{bmatrix} E_k^{-1} + \frac{1}{|E_{k+1}|}E_k^{-1}\underline{e}_k\underline{e}_k^T E_k^{-1} & -\frac{1}{|E_{k+1}|}E_k^{-1}\underline{e}_k \\ -\frac{1}{|E_{k+1}|}\underline{e}_k^T E_k^{-1} & \frac{1}{|E_{k+1}|} \end{bmatrix} \begin{bmatrix} \underline{1}_k \\ 1 \end{bmatrix} \\ &= \begin{bmatrix} E_k^{-1}\underline{1}_k + E_k^{-1}\underline{e}_k\frac{\underline{e}_k^T E_k^{-1}\underline{1}_k - 1}{|E_{k+1}|} \\ \frac{1 - \underline{e}_k^T E_k^{-1}\underline{1}_k}{|E_{k+1}|} \end{bmatrix}. \end{aligned}$$

Using the definitions of $|E_{k+1}|$ and α_k yields (42). □

Corollary 1 *For a given set of signal matrices $\{S_i\}_{i=1}^k$ with distinct outerproducts $\{S_i S_i^H\}_{i=1}^k$ assume that the k-element vector $\underline{x}^k = E_k^{-1}\underline{1}_k$ lies in $\mathbb{R}_+^k$. Let S_{k+1} be a signal such that $S_{k+1}S_{k+1}^H \notin \{S_i S_i^H\}_{i=1}^k$. Then the $(k+1)$-element vector $\underline{x}^{k+1} = E_{k+1}^{-1}\underline{1}_{k+1}$ is in $\mathbb{R}_+^{k+1}$ iff (a) $\alpha_k > 0$ and (b) $E_k^{-1}(\underline{1}_k - \alpha_k \underline{e}_k)$ is in $\mathbb{R}_+^k$. Furthermore (b) implies that $\alpha_k \leq 1$.*

Proof of Corollary 1

It only need be shown that condition (b) of the Corollary implies that $\alpha_k \leq 1$. If this condition holds then $\underline{e}_k^T E_k^{-1}(\underline{1}_k - \alpha_k \underline{e}_k) \geq 0$ as $\underline{e}_k$ has non-negative elements. Therefore

$$\alpha_k \underline{e}_k^T E_k^{-1} \underline{e}_k \leq \underline{e}_k^T E_k^{-1} \underline{1}_k. \tag{44}$$

By definition (43) $\alpha_k(1 - \underline{e}_k^T E_k^{-1}\underline{e}_k) = 1 - \underline{e}_k^T E_k^{-1}\underline{1}_k$. Adding this latter equation to the inequality (44) gives $\alpha_k \leq 1$. □

Corollary 1 motivates an iterative procedure, specified in Hero and Marzetta [13], for generating an monotonically improving sequence of constellations lying in the feasibility set $\tilde{\mathcal{S}}_{\text{peak}}^k$ defined in Proposition 4. At iteration k assume that a k-dimensional peak constrained constellation $S_1, \ldots, S_k$ has been constructed whose optimal distribution $\underline{P}_k^*$ attaining the minimum in (18) is an equalizer distribution: $\underline{P}_k^* = cE_k^{-1}\underline{1}_k$. Consider adding a candidate signal S_{k+1}, with $S_{k+1}S_{k+1}^H$ distinct from $\{S_i S_i^H\}_{i=1}^k$, having associated dissimilarity vector $\underline{e}_k$ given by (40). The two conditions (a) and (b) of the corollary may be used to select S_{k+1} to ensure that the optimal distribution $\underline{P}_{k+1}^*$ for the updated constellation $S_1, \ldots, S_{k+1}$ is also an equalizer distribution.

6.3 Feasibility via Polyhedral Inscribed Cones

Here we specify a simpler positive cone which is inscribed inside cone(E_k) and is generated by the positive definite $k \times k$ matrix

$$F_k(\delta) \stackrel{\text{def}}{=} I_k(1-\delta) + \delta \underline{1}_k \underline{1}_k^T, \tag{45}$$

where, $\delta \in [0, 1)$. We recall the Sherman-Morrissey-Woodbury identity for the inverse

$$F_k^{-1}(\delta) = \frac{1}{1-\delta}\left[I_k - \frac{\delta}{1+(k-1)\delta}\underline{1}_k \underline{1}_k^T\right]. \tag{46}$$

The set cone$(F_k(\beta))$ is centered along the positive ray $\{c\underline{1}_k : c > 0\}$. The set cone$(F_k(\beta))$ is equal to the positive orthant $\mathbb{R}_+^k$ for $\beta = 0$ while it approaches the aforementioned positive ray as β approaches 1. Furthermore cone$(F_k(\beta))$ is decreasing in β in the sense that cone$(F_k(\beta)) \subset$ cone$(F_k(\beta'))$ for $\beta \geq \beta'$.

Define the maximum, minimum, mean, and rms values of a real vector $\underline{a} = [a_1, \ldots, a_k]^T$

$$\min(\underline{a}) = \min_{1 \le i \le k} a_i, \quad \max(\underline{a}) = \max_{1 \le i \le k} a_i, \quad \text{avg}(\underline{a}) = \underline{1}_k^T \underline{a}/k, \quad \text{rms}(\underline{a}) = \sqrt{\underline{a}^T \underline{a}/k}.$$

Lemma 4 *Let $\{S_i\}_{i=1}^k$ be a set of signal matrices with distinct outerproducts $\{S_i S_i^H\}_{i=1}^k$ such that $E_k^{-1}\underline{1}_k \in \mathbb{R}_+^k$. Assume that $\text{cone}(F_k(\beta)) \subset \text{cone}(E_k)$ for $0 \le \beta < 1$. Let S_{k+1} be a signal such that $S_{k+1}S_{k+1}^H \notin \{S_i S_i^H\}_{i=1}^k$. Then the $(k+1)$-element vector $\underline{x}^{k+1} = E_{k+1}^{-1}\underline{1}_{k+1}$ lies in $\mathbb{R}_+^{k+1}$ if*

$$(a) \quad 0 < \alpha_k \le 1, \quad \text{and} \quad (b) \quad \max(\underline{e}_k) + \tfrac{k\beta}{1-\beta}\left[\max(\underline{e}_k) - \text{avg}(\underline{e}_k)\right] < 1 \tag{47}$$

where $\underline{e}_k$ and α_k are defined in (40) and (43).

An alternative form for (b) in Lemma 4 is:

$$(1 - \max(\underline{e}_k))/(1 - \text{avg}(\underline{e}_k)) > \theta(\beta), \tag{48}$$

where $\theta(z) \stackrel{\text{def}}{=} kz/(1 + (k-1)z) \in [0,1)$ is monotonic increasing over $z \in [0,1)$ and, as by Lemma 2 $\text{avg}(\underline{e}_k) < 1$, under condition (a): $1 - \alpha_k \text{avg}(\underline{e}_k) \ge 1 - \text{avg}(\underline{e}_k) > 0$.

Interpretation of the sufficient conditions in Lemma 4 is similar to the interpretation of Lemma 3. Condition (a) is equivalent to

$$\underline{e}_k^T E_k^{-1} \underline{e}_k \le \underline{1}_k^T E_k^{-1} \underline{e}_k < 1. \tag{49}$$

If $\underline{e}_k$ lies close to the column span of E_k then $1 - \underline{e}_k^T E_k^{-1}\underline{e}_k$ is small and condition (a) restricts the inner product of $\underline{e}_k$ and the previous un-normalized probability vector $\underline{x}^k = E_k^{-1}\underline{1}_k$ to a narrow range near 1. Thus we can expect that constraint (a) will become active only for densely packed signal constellations (large $K/(TM)$). When α_k is close to one, which occurs for the case of small values of $N \max_i D(S_i|S_{k+1})$ (low SNR), condition (b) places restrictions on the elements of feasible vectors $\underline{e}_k$ to ensure low variation about the mean value $\text{avg}(\underline{e}_k)$. This can be ensured if S_{k+1} can be selected to minimize the variation of its pairwise dissimilarities $\{D(S_i\|S_{k+1})\}_{i=1}^k$.

Proof of Lemma 4

Condition (a) of Lemma 4 obviously implies condition (a) of Corollary 1. Under the assumption $\text{cone}(F_k(\beta)) \subset \text{cone}(E_k)$ we will show conditions that (a) and (b) of Lemma 4 jointly imply condition (b) of Corollary 1. First, observe that $E_k^{-1}(\underline{1}_k - \alpha_k \underline{e}_k)$ is in $\mathbb{R}_+^k$ iff $\underline{1}_k - \alpha_k \underline{e}_k \in \text{cone}(E_k)$. Second, since $\text{cone}(E_k)$ contains $\text{cone}(F_k(\beta))$ it will suffice to show that $\underline{1}_k - \alpha_k \underline{e}_k \in \text{cone}(F_k(\beta))$ or, equivalently, $[F_k(\beta)]^{-1}(\underline{1}_k - \alpha_k \underline{e}_k)$

is in $\mathbb{R}_+^k$, under the conditions of Lemma 4. Applying the Sherman-Morrison-Woodbury identity (46) to $[F_k(\beta)]^{-1}$

$$[F_k(\beta)]^{-1}(\underline{1}_k - \alpha_k \underline{e}_k) = \frac{1}{1-\beta}\left[\underline{1}_k - \alpha_k \underline{e}_k - \frac{k\beta}{1+(k-1)\beta}(1-\alpha_k \text{avg}(\underline{e}_k))\underline{1}_k\right]. \quad (50)$$

The minimum element of the vector (50) is

$$\begin{aligned}\min\left([F_k(\beta)]^{-1}(\underline{1}_k - \alpha_k \underline{e}_k)\right) &= \frac{1}{1-\beta}\left[1 - \alpha_k \max(\underline{e}_k) - \frac{k\beta}{1+(k-1)\beta}(1-\alpha_k \text{avg}(\underline{e}_k))\right] \\ &= \frac{1}{1+(k-1)\beta}\left[1 - \alpha_k\left(\max(\underline{e}_k) + \frac{k\beta}{1-\beta}\left[\max(\underline{e}_k) - \text{avg}(\underline{e}_k)\right]\right)\right]\end{aligned}$$

which is positive under conditions (a) and (b) of the Lemma. Thus $[F_k(\beta)]^{-1}(\underline{1}_k - \alpha_k \underline{e}_k)$ is in the positive orthant which establishes condition (b) of Corollary 1. □

Since E_k is positive definite and cone$(F_k(\beta))$ and cone(E_k) are convex, $\underline{1}_k$ does not lie on the boundary of cone(E_k) and there always exists a value $\beta_o > 0$ such that cone$(F_k(\beta)) \subset$ cone(E_k) for all $\beta \in [\beta_o, 1)$. As the set cone$(F_k(\beta))$ is monotone decreasing in β the largest possible inscribed cone is obtained by using the minimum possible value of β. This gives the least restrictive sufficient condition (b) in (47). The form of this minimum β is specified for $k \geq 2$ in the Lemma below.

Lemma 5 *For a given set of signal matrices $\{S_i\}_{i=1}^k$ with distinct outerproducts $\{S_i S_i^H\}_{i=1}^k$ assume that the k-element vector $\underline{x}^k = E_k^{-1}\underline{1}_k$ lies in $\mathbb{R}_+^k$. For $k \geq 2$ the minimum value of $\beta \in [0,1)$ ensuring* cone$(F_k(\beta)) \subset$ cone(E_k) *is the value β_k given by*

$$\beta_k = \begin{cases} \frac{(k-1)-k\sqrt{(k-1)\gamma_{\min}(1-\gamma_{\min})}}{(k-1)(1-k\gamma_{\min})}, & \gamma_{\min} \neq \frac{1}{k} \\ \frac{1}{2} - \frac{\gamma_{\min}}{2}\,\frac{k}{k-1} & \gamma_{\min} = \frac{1}{k} \end{cases} \quad (51)$$

where $\gamma_{\min} = \min_i \gamma_k(i)$, $0 \leq \gamma_{\min} \leq 1/k$, and

$$\gamma_k(i) = \frac{1}{k}\underline{1}_k^T[I_k - \Pi_{E_k(-i)}]\underline{1}_k \quad (52)$$

and $\Pi_{E_k(-i)}$ is the orthogonal projector defined in (36).

Proof of Lemma 5

Fix a value $c > 0$. Let b be the distance between the vector $c\underline{1}_k$ and the closest face of cone(E_k) (See Fig. 2). As each face is contained in the column span of $E_k(-i))$ and as cone(E_k) is convex and contains $c\underline{1}_k$, b is specified by the projection theorem

$$b^2 = c^2 \min_i \left\{\underline{1}_k^T[I_k - \Pi_{E_k(-i)}]\underline{1}_k\right\} = c^2 k\,\gamma_{\min}, \quad (53)$$

Since $I_k - \Pi_{E_k(-i)}$ is an idempotent matrix $\gamma_k(i) = \|[I_k - \Pi_{E_k(-i)}]\underline{1}_k\|^2/k \geq 0$. As the elements of E_k are non-negative the projection error norm $\underline{1}_k^T[I_k - \Pi_{E_k(-i)}]\underline{1}_k$ is maximum when cone(E_k) is largest, which occurs for $E_k = I_k$. In this case $I_k - \Pi_{E_k(-i)} = \text{diag}(0, \ldots, 0, 1, 0, \ldots, 0)$ and $\gamma_k(i) = 1/k$.

Next let $d(\beta)$ be the maximum distance from $c\underline{1}_k$ to cone$(F_k(\beta))$ (See Figure 2). As the extremals of cone$(F_k(\beta))$ are symmetric about the ray $\{a\underline{1}_k : a > 0\}$ each extremal has a point at identical maximum distance from $c\underline{1}_k$. Consider the extremal $\{a\underline{f}_1 : a > 0\}$ where $\underline{f}_1$ is the first column of $F_k(\beta)$. We thus have

$$\begin{aligned} d^2(\beta) &= c^2 \left\| \underline{1}_k - \frac{\underline{1}_k^T \underline{f}_1}{\|\underline{f}_1\|^2} \underline{f}_1 \right\|^2 \\ &= c^2 k \left[1 - \frac{(1 + (k-1)\beta)^2}{k(1 + (k-1)\beta^2)} \right]. \end{aligned} \tag{54}$$

Evidently cone$(F_k(\beta)) \subset$ cone(E_k) as long as $d^2(\beta) \leq b^2$. Let β_k be the minimum corresponding value of $\beta \in [0, 1)$ for which $d^2(\beta_k) = b^2$. Equating (53) and (54) yields the following quadratic equation for β_k

$$(k-1)(\zeta(k-1) - 1)\beta_k^2 + 2(k-1)\zeta\beta_k + \zeta - 1 = 0$$

where $\zeta = (1 + \gamma_{\min}/(1 - \gamma_{\min}))/k$. For $\gamma_{\min} \in [0, 1/k)$ the quadratic equation has two non-negative solutions only one of which is contained in $[0, 1)$. For $\gamma_{\min} = 1/k$, $\zeta = 1/(k-1)$ and there is one solution in $[0, 1)$. In both cases this solution is given by (51). □

7 Low Rate Optimality of Unitary Orthogonal Constellations

When the number of signal matrices K to be considered is sufficiently small significant simplification of the optimization (17) is possible. In particular, one obtains optimality of a set of scaled mutually orthogonal unitary signal matrices and a simple form for $\tilde{R}_o(K)$.

The first result specifies the solution to optimization of the dissimilarity measure $D(S_i\|S_j)$ defined in (10).

For given η, T and M define the integer M_o

$$M_o = \text{argmax}_{m \in \{1, \ldots, M\}} \left\{ m \ln \frac{(1 + \eta TM/(2m))^2}{1 + \eta TM/m} \right\}. \tag{55}$$

We will see below that under some conditions M_o is the rank of the signal matrices S_i in the optimal K-dimensional constellation.

Proposition 8 *Let $2M \leq T$. Then*

$$D_{\max} \stackrel{\text{def}}{=} \max_{S_1, S_2 \in \mathcal{S}^K_{\text{peak}}} D(S_1 \| S_2) = M_o \ln \frac{(1 + \eta TM/(2M_o))^2}{1 + \eta TM/M_o}. \tag{56}$$

Furthermore, the optimal signal matrices which attain $D_{\max}$ can be taken as scaled rank M_o mutually orthogonal unitary $T \times M$ matrices of the form

$$S_1 = \sqrt{TM/M_o}\, \Phi_1, \qquad S_2 = \sqrt{TM/M_o}\, \Phi_2$$

where, for $j = 1, 2$,

$$\Phi_j^H \Phi_j = I_{M_o}, \quad \text{and} \quad \Phi_i^H \Phi_j = 0, \;\; i \neq j.$$

The assumption $2M \leq T$ is critical and ensures that the singular vectors of S_1 and S_2 can be chosen as mutually orthogonal for any set of singular values.

The rank M_o of the optimal matrices S_1 and S_2 increases from 1 to M as the SNR parameter ηTM increases from 0 to ∞ (see Fig. 3). Numerical evaluation has shown that the functional relationship between M_o and SNR is well approximated by the relation

$$M_o \approx \max\left(1, \lfloor a\eta TM + b + 0.48 \rfloor\right)$$

where a, b are the slope and intercept of the least squares linear fit to the function $y(x) = \text{argmax}_{m=1,2,\ldots} m \ln[(1 + x/(2m))^2/(1 + x/m)]$. The approximation is a lower bound and underestimates the exact value of M_o, given by (55), by at most 1 over less than 0.5% of the SNR range shown in Fig. 3 ($0 < \eta TM \leq 120$). If the SNR is sufficiently large, e.g. (from Fig. 3) $\eta TM \geq 17$ for $M = 6$ and $T \geq 12$, $M_o = M$ and the optimal signal matrices utilize all M transmit antennas. On the other hand for small SNR, i.e. (from Fig. 3) $\eta TM < 4$, $M_o = 1$ and the optimal signal matrices apply all available transmit power to a single antenna element over the coherent fade interval T.

Proof of Proposition 8

Let S_1 and S_2 have the singular value decompositions

$$S_1 = V_1 \Lambda_1 U_1^H, \qquad S_2 = V_2 \Lambda_2 U_2^H, \tag{57}$$

where V_1, V_2 are $T \times M$ unitaries, i.e. $V_1^H V_1 = V_2^H V_2 = I_M$, U_1, U_2 are $M \times M$ unitaries, and Λ_1, Λ_2 are $M \times M$ real diagonal matrices of singular values $\{\lambda_{1i}\}_{i=1}^M$ and $\{\lambda_{2i}\}_{i=1}^M$, respectively.

The function $D(S_1\|S_2)$ of $S_1S_1^H$ and $S_2S_2^H$ only depends on S_1 and S_2 through $V_1, \Lambda_1, V_2, \Lambda_2$. Thus unitaries U_1 and U_2 can be arbitrarily chosen without affecting $D(S_1\|S_2)$ and hence we can choose them as I_M. We denote this functional dependence by writing $D(S_1\|S_2) = D(V_1, \Lambda_1\|V_2, \Lambda_2)$. Therefore

$$\max_{S_1,S_2} D(S_1\|S_2) = \max_{\Lambda_1,\Lambda_2} \max_{V_1,V_2} D(V_1, \Lambda_1\|V_2, \Lambda_2). \tag{58}$$

Using the form for D given in Lemma 6 (see the Appendix)

$$\begin{aligned} D(V_1, \Lambda_1\|V_2, \Lambda_2) &= \ln \frac{\left|I_M + \frac{\eta}{2} S_1^H S_1\right| \left|I_M + \frac{\eta}{2} S_2^H S_2\right|}{\sqrt{\left|I_M + \eta S_1^H S_1\right| \left|I_M + \eta S_2^H S_2\right|}} + \ln \left|I_M - \kappa^H \kappa\right| \\ &= \sum_{i=1}^{M} \ln \frac{(1+\frac{\eta}{2}\lambda_{1i}^2)(1+\frac{\eta}{2}\lambda_{2i}^2)}{\sqrt{(1+\eta\lambda_{1i}^2)(1+\eta\lambda_{2i}^2)}} + \ln \left|I_M - \kappa^H \kappa\right| \end{aligned} \tag{59}$$

For fixed arbitrary Λ_1, Λ_2 first consider the inner maximization in (58), i.e. maximization over V_1, V_2 of the right hand side of (59). Observe that the first term in (59) depends only on the singular values Λ_1, Λ_2 and not on V_1, V_2. Observe that as $S_1^H S_2 = \Lambda_1 V_1^H V_2 \Lambda_2$, for any Λ_1, Λ_2 the product $S_1^H S_2$ can be forced to zero by selecting matrices V_1 and V_2 to be mutually orthogonal – possible under the assumption $2M \leq T$. Recall that $\kappa = \tilde{S}_2^H \tilde{S}_1$ where $\tilde{S}_1$ and $\tilde{S}_2$, defined in Lemma 6, have the same column spaces as S_1 and S_2. Therefore, when V_1 and V_2 are mutually orthogonal $\kappa = 0$ and the second term in (59) is equal to its maximum value (0). This establishes that the optimum signal matrices S_1 and S_2 have orthogonal singular vectors V_1 and V_2 and that

$$\max_{\Lambda_1,\Lambda_2} \max_{V_1,V_2} D(V_1, \Lambda_1\|V_2, \Lambda_2) = \max_{\{\lambda_{1i}\}} \sum_{i=1}^{M} \ln \frac{1+\frac{\eta}{2}\lambda_{1i}^2}{\sqrt{1+\eta\lambda_{1i}^2}} + \max_{\{\lambda_{2i}\}} \sum_{i=1}^{M} \ln \frac{1+\frac{\eta}{2}\lambda_{2i}^2}{\sqrt{1+\eta\lambda_{2i}^2}}.$$

Due to the assumed peak power constraint each of the above maximizations on the right hand side are performed over the inequality constraint sets $\left\{\lambda_{j1}, \ldots, \lambda_{jM} : \sum_{i=1}^{M} \lambda_{ji}^2 \leq TM\right\}$, $j = 1, 2$. As these maximizations are of identical form,

$$\max_{\Lambda_1,\Lambda_2} \max_{V_1,V_2} D(V_1, \Lambda_1\|V_2, \Lambda_2) = 2 \max_{\{\lambda_i\}} \sum_{i=1}^{M} \ln \frac{1+\frac{\eta}{2}\lambda_i^2}{\sqrt{1+\eta\lambda_i^2}} = \max_{\{\lambda_i\}} \sum_{i=1}^{M} \ln \frac{(1+\frac{\eta}{2}\lambda_i^2)^2}{1+\eta\lambda_i^2} \tag{60}$$

where the maximization on the right hand side is subject to $\sum_{i=1}^{M} \lambda_i^2 \leq TM$. Observe that as each summand is monotone increasing in λ_i^2 the inequality constraint in the maximization is always met with equality. Consider the Lagrangian

$$J(\lambda_1, \ldots, \lambda_M) = \sum_{i=1}^{M} \left(2\ln(1+\tfrac{\eta}{2}\lambda_i^2) - \ln(1+\eta\lambda_i^2)\right) - \alpha \sum_{i=1}^{M} \lambda_i^2$$

where $\alpha \geq 0$ is an undetermined multiplier. For a suitable value of α, the maximum of J over unconstrained $\{\lambda_i\}$ is identical to the maximum of the right hand side of (60) over power constrained $\{\lambda_i\}$. The derivative

of J with respect to λ_i is

$$\frac{\partial J}{\partial \lambda_i} = -\alpha\eta^2 \, \lambda_i \frac{(\lambda_i^2)^2 + 2\lambda_i^2 \, v/\eta + 2/\eta^2}{(1+\eta\lambda_i^2)(1+\frac{\eta}{2}\lambda_i^2)}$$

where $v = \frac{3\alpha-\eta}{2\alpha}$. Therefore, for each λ_i, there are five possible stationary points of J given by either $\lambda_i = 0$ or by $\overset{+}{-}\sqrt{\lambda_i^2}$ where λ_i^2 is one of the roots

$$\lambda_i^2 = (-v \overset{+}{-} \sqrt{v^2-2})/\eta, \quad i = 1, \ldots, M. \tag{61}$$

With respect to (61) there are three cases which must be considered. For $v \geq \sqrt{2}$ both roots are real negative and for $-\sqrt{2} < v < \sqrt{2}$ both roots are complex. For $v \leq -\sqrt{2}$ both roots are real positive. Thus only the latter case is relevant. It can be verified that, when λ_i^2 is equal to one of these two positive roots, the second derivative of J is $\partial^2 J/\partial\lambda_i^2 = -4\alpha\eta^2\lambda_i^2 \frac{\lambda_i^2 + v/\eta}{(1+\eta\lambda_i)(1+\frac{\eta}{2}\lambda_i)}$ which is negative only for the root $\lambda_i^2 = (-v + \sqrt{v^2-2})/\eta$. Therefore, we can restrict attention to a candidate maximizer $\{\lambda_1, \ldots, \lambda_M\}$ of J for which each λ_i^2 can take on either of two values: zero or $c = (-v + \sqrt{v^2-2})/\eta$. Let $m \in \{1, \ldots, M\}$ denote the number of nonzero valued λ_i for one of these candidate maximizers. Then, invoking the constraint $\sum_{i=1}^M \lambda_i^2 = TM$ and noting that the ordering of the indices of λ_i^2 is irrelevant, any candidate maximizer of J can be put in the form

$$\lambda_i^2 = \begin{cases} TM/m, & i = 1, \ldots, m \\ 0, & i = m+1, \ldots, M \end{cases}, \tag{62}$$

and now m is the only remaining free parameter. Substituting this expression for $\nu_i^2 = \lambda_i^2$ into (60) we obtain the expression (56) with M_o equal to the optimal value of m given by (55). Using the SVD representation (57) with $U_1 = U_2 = I_M$ it is easily seen that the optimal signal matrices, corresponding to (62) with $m = M_o$, have the form $S_j = V_j\Lambda_j = \sqrt{TM/M_o}\ \Phi_j, \qquad j = 1, 2$, where the Φ_j's are rank M_o mutually orthogonal rectangular unitaries as defined in the statement of the Proposition. □

Proposition 8 implies that for low symbol rate, the distance-optimal signal constellations are constellations of scaled mutually orthogonal unitary matrices of rank $M_o \leq M$.

Corollary 2 *Let $2M \leq T$ and let M_o be as defined in (55). Assume $M_o \leq \min\{M, T/K\}$ and define the set of signal matrices $\{S_i^*\}_{i=1}^K$ by $S_i^* = \sqrt{TM/M_o}\,\Phi_i$ where $\{\Phi_i\}_{i=1}^K$ are a set of $T \times M$ mutually orthogonal rectangular unitary matrices of rank M_o ($\Phi_i^H\Phi_i = I_{M_o}$ and $\Phi_i^H\Phi_j = 0$, $i \neq j$). This set of signal matrices are equidistant in the sense $D(S_i^*\|S_j^*) = D_{\max} = D_{\min}$, $i \neq j$, and they attain the maximum possible value of $D_{\min} = \min_{l\neq m} D(S_l\|S_m)$ over all signal sets $\{S_i\}_{i=1}^K \in \mathcal{S}_{\text{peak}}^K$ of dimension K.*

Proof of Corollary 2

First observe that if $M_o \leq \min\{M, T/K\}$ then such an orthogonal unitary set $\{\Phi_i\}_{i=1}^K$ exists, e.g. the set of K mutually orthogonal permutation matrices of dimension $T \times M$ whose columns are formed from K disjoint M_o-dimensional subsets of the columns of I_T. The set $\{S_i^*\}_{i=1}^K$ satisfies the peak power constraint $\|S_i^*\|^2 \leq TM$, $i = 1, \ldots, K$, with equality and, by Proposition 8, it has the property that $D(S_i^* \| S_j^*) = D_{\max} \stackrel{\text{def}}{=} \max_{S_1, S_2} D(S_1 \| S_2)$, $i \neq j$. Let $\{S_i\}_{i=1}^K$ be any set of K signal matrices in $\mathbb{C}^{T \times M}$ satisfying the same power constraint. Then $\min_{i \neq j} D(S_i \| S_j) \leq D_{\max} = \min_{i \neq j} D(S_i^* \| S_j^*)$ and thus $\{S_i^*\}_{i=1}^K$ maximizes $D_{\min}$. □

It can be easily verified that the mutually orthogonal unitary constellation of Corollary 2 satisfies the sufficient conditions (a) and (b) of Lemma 4 when $\tilde{\mathcal{S}}_{\text{peak}}^K$ is not empty. For this constellation $E_K = F_K(\delta)$, $\delta = e^{-ND_{\max}} < 1$, and

$$\begin{aligned} \alpha_K &= (1 - \underline{1}_K^T [F_K(\delta)]^{-1} \underline{1}_K \delta) / (1 - \underline{1}_K [F_K(\delta)]^{-1} \underline{1}_K \delta^2) \\ &= (1 - \theta)/(1 - \delta\theta). \end{aligned}$$

where $\theta = \theta(\delta) \in [0, 1)$ is defined below (48) with $k = K$. Therefore $0 < \alpha_K \leq 1$. Furthermore, as $\max(\underline{e}_K) = \text{avg}(\underline{e}_K) = \delta$, condition (b) reads: $\alpha_K \delta < 1$, which is satisfied regardless of the value of β.

The final result of this section is an expression for the cut-off rate.

Proposition 9 *Let $2M \leq T$ and let M_o be as defined in (55). Suppose that $M_o \leq \min\{M, T/K\}$. Then the peak constrained K dimensional cut-off rate (12) is*

$$\tilde{R}_o(K) = \ln\left(\frac{K}{1 + (K-1)e^{-ND_{\max}}}\right)$$

and $D_{\max}$ is given by (56). Furthermore, an optimal constellation attaining $\tilde{R}_o(K)$ is the set of K rank M_o mutually orthogonal matrices specified in Corollary 2 and the optimal probability assignment is uniform: $P_i^* = 1/K$, $i = 1, \ldots, K$.

Any unitary transformation on the columns (spatial coordinates) of a set of signal matrices produces a set of signal matrices with identical $D_{\min}$. In particular, any set of K mutually orthogonal $T \times M_o$ permutation matrices, specified in the course of proof of Corollary 2, has optimal distance properties. This simple set of signal matrices corresponds to transmitting energy on a single antenna element at a time, among a total of $M_o \leq M$ elements, in each of the available T time slots. Since $\tilde{R}_o(K)$ is increasing in K the maximum cut-off rate achievable using these mutually orthogonal unitary matrices is obtained by using the maximum

possible number of them: $K = \lfloor T/M_o \rfloor$. Observe that the resulting optimal constellation may correspond to a code of quite low symbol rate, e.g. for $M_o = M = T/2$ the symbol rate is only 1 bit-per-symbol.

It is noteworthy that the optimal peak constrained signal constellation specified by Proposition 9 does not include the zero valued signal matrix $S_i = O$. Including zero in the signal constellation would allow signalling using on-off keying. On-off keying is often proposed for average power constrained signalling over low SNR channels since it permits energy discrimination at the receiver. As contrasted with on-off keying all signals in the optimal peak constrained signal set have equal power. We conjecture that the zero signal would result from replacing the peak power constraint with an average power constraint in Proposition 9.

Proof of Proposition 9

Define $\delta = e^{-ND_{\max}}$ which is strictly less than one under the hypotheses of the Proposition. For any set of signal matrices $\{S_i\}_{i=1}^K$ satisfying the assumptions of Proposition 9, Proposition 8 asserts that $D(S_i\|S_j) \leq D_{\max}$, $i \neq j$, with equality when $\{S_i\}_{i=1}^K$ consists of the specified mutually orthogonal matrices. Using this inequality and the fact $D(S_i\|S_j) = 0$, $i = j$,

$$\begin{aligned} \min_{\{S_i\}_{i=1}^K \in \mathcal{S}_{\text{peak}}^K} \sum_{i=1}^K P_i \sum_{j=1}^K P_j \, e^{-ND(S_i\|S_j)} &= e^{-ND_{\max}} \sum_{i \neq j} P_i P_j + \sum_{i=1}^K P_i^2 \\ &= \underline{P}_K^T F_K(\delta) \underline{P}_K. \end{aligned} \tag{63}$$

Now, using the identity (46), $F_K(\delta)\underline{x} = \underline{1}_K$ has a unique solution in the positive orthant

$$\begin{aligned} \underline{x} &= [F_K^{-1}(\delta)]^{-1} \underline{1}_K \\ &= \frac{1}{1-\delta} \left(1 - \frac{K\delta}{1 + (K-1)\delta} \right) \underline{1}_K \\ &= \frac{1}{(1 + (K-1)\delta)} \underline{1}_K. \end{aligned}$$

Thus the minimizer $\underline{P}_K^*$ of $\underline{P}_K^T F_K(\delta) \underline{P}_K$ is the uniform distribution $\underline{P}_K = \underline{1}_K / K$. Substitution of this solution back into (63) establishes that

$$\min_{\{P_i\}_{i=1}^K \in \mathcal{P}^K} \min_{\{,S_i\}_{i=1}^K \in \mathcal{S}_{\text{peak}}^K} \sum_{i=1}^K P_i \sum_{j=1}^K P_j \, e^{-ND(S_i\|S_j)} = \frac{1}{K^2} \underline{1}_K^T F_K(\delta) \underline{1}_K = \frac{(1-\delta) + \delta K}{K}$$

which when substituted into (12) yields the expression given for $\tilde{R}_o(K)$ specified in Proposition 9. □

8 Conclusions

We have derived representations for the single user computational cut-off rate for space time coding under the Rayleigh quasi-static fading channel model under a peak transmitted power constraint. For finite dimensional

constellations the cut-off rate and the optimal signal distribution were specified as a solution to a quadratic optimization problem and it was shown that optimal constellations have codeword distributions which satisfy an equalization condition. This characterization of optimality motivated us to study properties of the set of feasible constellations which satisfy the equalization property. In particular, we showed that distance-optimal constellations are close to rate-optimal for large number of receive antennas. Easily verifiable necessary and sufficient conditions were given for validating that a given signal constellation lies in the feasible set. A recursive form was given for one of these conditions which may be useful for design of feasible constellations.

Acknowledgement

The authors would like to thank the reviewers and the associate editor for their thoughtful comments on this work.

9 Appendix: Properties of $D(S_1\|S_2)$

Here we collect various properties of the signal dissimilarity measure (10).

9.1 Alternative form for $D(S_1\|S_2)$

$D(S_1\|S_2)$ can be equivalently expressed in terms of a signal multiple correlation matrix κ.

Lemma 6 *The signal dissimilarity measure (10) for the spatio-temporal fading model (1) has the equivalent form*

$$D(S_1\|S_2) = \frac{1}{2}\ln\frac{\left|I_M+\frac{\eta}{2}S_1^HS_1\right|^2\left|I_M+\frac{\eta}{2}S_2^HS_2\right|^2}{\left|I_M+\eta S_1^HS_1\right|\left|I_M+\eta S_2^HS_2\right|}\left|I_M-\kappa^H\kappa\right|^2,$$

where κ is the $M\times M$ multiple correlation matrix

$$\kappa = \tilde{S}_2^H\tilde{S}_1,$$

$\tilde{S}_1$ and $\tilde{S}_2$ are the "prewhitened" signal matrices

$$\tilde{S}_1 = \sqrt{\tfrac{\eta}{2}}S_1[I_M+\tfrac{\eta}{2}S_1^HS_1]^{-\frac{1}{2}}, \qquad \tilde{S}_2 = \sqrt{\tfrac{\eta}{2}}S_2[I_M+\tfrac{\eta}{2}S_2^HS_2]^{-\frac{1}{2}},$$

and $A^{\frac{1}{2}}$ denotes the positive definite square root factor of positive definite square matrix A.

Proof of Lemma 6

Consider the quantity in the numerator of expression (10)

$$\begin{aligned}\left|I_T + \tfrac{\eta}{2}(S_1 S_1^H + S_2 S_2^H)\right| &= \left|I_T + \tfrac{\eta}{2}[S_1 \;\; S_2]\begin{bmatrix} S_1^H \\ S_2^H \end{bmatrix}\right| \\ &= \left|I_T + \tfrac{\eta}{2}\begin{bmatrix} S_1^H \\ S_2^H \end{bmatrix}[S_1 \;\; S_2]\right| \\ &= \left|\begin{bmatrix} I_M + \frac{\eta}{2}S_1^H S_1 & \eta S_1^H S_2 \\ \eta S_2^H S_1 & I_M + \frac{\eta}{2}S_2^H S_2 \end{bmatrix}\right|,\end{aligned}$$

where in the second line we have used the property $|I + AB| = |I + BA|$.

For A and C Hermitian positive definite matrices we recall the following determinant identity for block partitioned matrices

$$\left|\begin{bmatrix} A & B \\ B^H & C \end{bmatrix}\right| = |A| \;\; |C| \;\; \left|I - B^H A^{-1} B C^{-1}\right|.$$

Furthermore, denoting by $C^{\frac{1}{2}}$ the invertible square root factor of C, where $C = C^{\frac{H}{2}} C^{\frac{1}{2}}$,

$$\begin{aligned}\left|I - B^H A^{-1} B C^{-1}\right| &= \left|C^{\frac{H}{2}}(I - C^{-\frac{H}{2}} B^H A^{-\frac{1}{2}} A^{-\frac{H}{2}} B C^{-\frac{1}{2}}) C^{-\frac{H}{2}}\right| \\ &= \left|I - \kappa^H \kappa\right|\end{aligned}$$

where $\kappa = C^{-\frac{H}{2}} B^H A^{-\frac{1}{2}}$. Identifying $A = I_M + \frac{\eta}{2} S_1^H S_1$, $C = I_M + \frac{\eta}{2} S_2^H S_2$, and $B = S_1^H S_2$ finishes the proof of Lemma 6. □

9.2 Bounds on $D(S_1 \| S_2)$

Lemma 7 *The dissimilarity measure $D(S_1 \| S_2)$ satisfies*

$$D(S_1 \| S_2) = \frac{\eta^2}{8} \| S_1 S_1^H - S_2 S_2^H \|^2 + o(\eta^2)$$

where $o(\eta^2)$ is a non-negative function of η such that $\lim_{\eta \to 0} o(\eta^2)/\eta^2 = 0$.

Proof of Lemma 7

Let $\{\nu_{1i}\}_{i=1}^T$ and $\{\nu_{2i}\}_{i=1}^T$ denote the eigenvalues of $A = S_1 S_1^H$ and $B = S_2 S_2^H$ respectively and let $\{\sigma_i\}_{i=1}^T$ denote the eigenvalues of $\frac{1}{2}(A + B)$. Since these are Hermitian non-negative definite matrices these eigenvalues are real and positive. Thus the dissimilarity measure $D(S_1 \| S_2)$, given in (10), can be written as

$$D(S_1 \| S_2) = \tfrac{1}{2} \sum_{i=1}^{T} \left(2 \ln(1 + \eta \sigma_i) - \ln(1 + \eta \nu_{1i}) - \ln(1 + \eta \nu_{2i})\right).$$

Using the relation $\ln(1+a) = a - a^2/2 + o(a^2)$ and the fact that $\sum_{i=1}^{T}(2\sigma_i - \nu_{1i} - \nu_{2i}) = 0$

$$\begin{aligned} D(S_1\|S_2) &= \frac{\eta^2}{4}\sum_{i=1}^{T}\left(\nu_{1i}^2 + \nu_{2i}^2 - 2\sigma_i^2\right) + o(\eta^2) \\ &= \frac{\eta^2}{4}\left(\mathrm{tr}(AA^H) + \mathrm{tr}(BB^H) - 2\mathrm{tr}\left(\frac{(A+B)}{2}\frac{(A+B)^H}{2}\right)\right) + o(\eta^2) \\ &= \frac{\eta^2}{8}\mathrm{tr}\left((A-B)(A-B)^H\right) + o(\eta^2). \end{aligned}$$

Since $\mathrm{tr}((A-B)(A-B)^H) = \|A-B\|^2$ Lemma 7 is established. □

Lemma 8 *Let A and B be $T \times T$ complex matrices. Then the trace norm of $A - B$ satisfies*

$$\|A-B\|^2 = \mathrm{tr}((A-B)(A-B)^H) \geq \sum_{i=1}^{T}|a_{ii} - b_{ii}|^2, \tag{64}$$

where $\{a_{ii}\}_{i=1}^{T}$ and $\{b_{ii}\}_{i=1}^{T}$ are the diagonal elements of A and B, respectively. Equality occurs in (64) iff the trace norm of the anti-diagonal matrix $A - B - \mathrm{diag}(A-B)$ is equal to zero.

Proof of Lemma 8

Let $\Delta = A - B$ have diagonal elements $\{d_{ii}\}_{i=1}^{T}$ and consider the decomposition $\Delta = D + Z$ into diagonal matrix $D = \mathrm{diag}(d_{ii})$ and anti-diagonal matrix Z. Then

$$\|\Delta\|^2 = \|D\|^2 + \|Z\|^2 + \mathrm{tr}(DZ^H) + \mathrm{tr}(ZD^H).$$

As $\mathrm{tr}(DZ^H) = \mathrm{tr}(ZD^H) = 0$,

$$\|\Delta\|^2 \geq \|D\|^2 = \sum_{i=1}^{T} d_{ii}^2,$$

with equality iff $\|Z\| = 0$. □

The following establishes that it is safe to assume that the signal matrices in the optimal constellation have distinct outerproducts. While this result can also be obtained from the statistical invariance of the Rayleigh fading model, the lemma below is proven using a more elementary non-statistical argument.

Lemma 9 *Let the set of signal matrices $\{S_i\}_{i=1}^{K}$ have dissimilarity matrix E_K and assume that for some $i \neq j$: $S_iS_i^H = S_jS_j^H$. Then there exists a $K-1$ dimensional subset of $\{S_i\}_{i=1}^{K}$ having dissimilarity matrix E_{K-1} such that*

$$\min_{\underline{P}_K \in \mathcal{P}^K} \underline{P}_K^T E_K \underline{P}_K = \min_{\underline{P}_{K-1} \in \mathcal{P}^{K-1}} \underline{P}_{K-1}^T E_{K-1} \underline{P}_{K-1}$$

Proof of Lemma 9

Without loss of generality we can assume that $S_{K-1}S_{K-1}^H = S_K S_K^H$. As $D(S_i\|S_j)$ is a function of S_i and S_j only through $S_iS_i^H$ and $S_jS_j^H$, E_K takes the form

$$E_K = \begin{bmatrix} E_{K-2} & \underline{e} & \underline{e} \\ \underline{e}^T & 1 & 1 \\ \underline{e}^T & 1 & 1 \end{bmatrix}, \qquad ,$$

where $\underline{e}$ is a vector with entries $\exp(-ND(S_i\|S_K)) = \exp(-ND(S_i\|S_{K-1}))$, $i = 1, \ldots, K-2$. Let $\underline{P}_{K-2} = [P_1, \ldots, P_{K-2}]^T$ and $\tilde{Q}(\{P_i\}_{i=1}^K, \{S_i\}_{i=1}^K) = \underline{P}_K^T E_K \underline{P}_K$. Then using (9.2)

$$\begin{aligned} \tilde{Q}\left(P_1, \ldots, P_{K-1}, P_K, S_1, \ldots, S_{K-1}, S_K\right) &= (P_K + P_{K-1})^2 + 2[\underline{P}_{K-2}]^T \underline{e}(P_K + P_{K-1}) + [\underline{P}_{K-2}]^T E_{K-2}\underline{P}_{K-2} \\ &= \tilde{Q}\left(P_1, \ldots, P_{K-2}, (P_K + P_{K-1}), S_1, \ldots, S_{K-2}, S_{K-1}\right) \end{aligned}$$

Since

$$\min_{\underline{P}_K \in \mathcal{P}^K} \tilde{Q}\left(P_1, \ldots, P_{K-2}, (P_K + P_{K-1}), S_1, \ldots, S_{K-2}, S_{K-1}\right) = \min_{\underline{P}_{K-1} \in \mathcal{P}^{K-1}} \tilde{Q}\left(P_1, \ldots, P_{K-1}, S_1, \ldots, S_{K-1}\right)$$

the Lemma follows. □

References

[1] I. Abou-Faycal and B. M. Hochwald, "Coding requirements for multiple-antenna channels with unknown Rayleigh fading," Technical report, Bell Laboratories, Lucent Technologies, Murray Hill, NJ, 1999.

[2] A. Berman and R. J. Plemmons, *Nonnegative matrices in the mathematical sciences*, SIAM, Philadelphia PA, 1994.

[3] C. Berrou, A. Glavieux, and P. Thitimajshima, "Near Shannon limit error-correcting coding and decoding: turbo codes," in *Proc. Int. Conf. on Commun.*, pp. 1064–1070, 1993.

[4] E. Biglieri, J. Proakis, and S. S. (Shitz), "Fading channels: information-theoretic and communications aspects," *IEEE Trans. on Inform. Theory*, vol. 44, pp. 2619–2692, Oct., 1998.

[5] R. Blahut, *Principles and Practice of Information Theory*, Addison-Wesley, 1987.

[6] T. Cover and J. Thomas, *Elements of Information Theory*, Wiley, New York, 1987.

[7] G. J. Foschini, "Layered space-time architectures for wireless communication in a fading environment when using multi-element antennas," *Bell Labs Tech. Journ.*, vol. 1, no. 6, pp. 41–49, 1996.

[8] G. J. Foschini and M. J. Gans, "On limits of wireless communication in a fading environment when using multiple antennas," *Wireless Personal Commun.*, vol. 6, no. 3, pp. 311–335, 1998.

[9] R. G. Gallager, "A simple derivation of the coding theorem and some applications," *IEEE Trans. on Inform. Theory*, vol. IT-11, pp. 3–18, Jan. 1965.

[10] R. G. Gallager, *Information Theory and Reliable Communication*, Wiley, 1968.

[11] J. Hagenauer, "Iterative decoding of binary block and convolutional codes," *IEEE Trans. on Inform. Theory*, vol. IT-42, no. 2, pp. 429–445, Mar. 1996.

[12] B. Hassibi and B. M. Hochwald, "How much training is needed in multiple-antenna wireless links," Technical report, Lucent Bell Laboratories Technical Memo, 2000.

[13] A. O. Hero and T. L. Marzetta, "On computational cut-off rate for space time coding," Technical report, Bell Laboratories, Lucent Technologies, Murray Hill, NJ, 2000. `http://mars.bell-labs.com/cm/ms/what/mars/index.html`.

[14] B. M. Hochwald and T. L. Marzetta, "Unitary space-time modulation for multiple-antenna communications in Rayleigh flat fading," Technical report, Bell Laboratories, Lucent Technologies, Murray Hill, NJ, 1999.

[15] B. M. Hochwald and T. L. Marzetta, "Unitary space-time modulation for multiple-antenna communication," *IEEE Trans. on Inform. Theory*, vol. IT-46, pp. 543–564, Mar. 2000.

[16] B. M. Hochwald, T. L. Marzetta, T. J. Richardson, W. Sweldens, and R. Urbanke, "Systematic design of unitary space-time constellations," *IEEE Trans. on Inform. Theory*, vol. IT-46, pp. 1962–1973, Sept. 2000.

[17] I. A. Ibragimov and R. Z. Has'minskii, *Statistical estimation: Asymptotic theory*, Springer-Verlag, New York, 1981.

[18] S. Kallel and K. Li, "Biderectional sequential decoding," *IEEE Trans. on Inform. Theory*, vol. IT-43, pp. 1319–1326, July 1997.

[19] S. M. Kay, *Modern Spectral Estimation: Theory and Application*, Prentice-Hall, Englewood-Cliffs N.J., 1988.

[20] T. L. Marzetta and B. M. Hochwald, "Capacity of a mobile multiple-antenna communication link in Rayleigh fading," *IEEE Trans. on Inform. Theory*, vol. IT-45, pp. 139–158, Jan. 1999.

[21] J. L. Massey, "Variable length codes and the Fano metric," *IEEE Trans. on Inform. Theory*, vol. IT-19, no. 1, pp. 196–198, 1972.

[22] J. L. Massey, "Coding and modulation in digital communications," in *Proc. 1974 Int. Zurich Sem, Digital Communic*, pp. E2.1–E2.4, Mar. 1974.

[23] J. J. Moré and S. J. Wright, *Optimization software guide*, Frontiers in Applied Mathematics Series, SIAM, Philadelphia PA, 1993.

[24] P. Narayan and D. L. Snyder, "The two-user cutoff rate region for an asynchronous and a synchronous multiple access channel are the same," *IEEE Trans. on Inform. Theory*, vol. IT-27, pp. 414–419, July 1981.

[25] P. Narayan and D. L. Snyder, "Signal set design for band-limited memoryless multiple access channels with soft decision demodulation," *IEEE Trans. on Inform. Theory*, vol. IT-33, pp. 539–556, July 1987.

[26] T. Rockafellar, *Convex Analysis*, Princeton Univ. Press, Princeton, NJ, 1970.

[27] J. E. Savage, "Sequential decoding – the computation problem," *Bell Syst. Tech. Journ.*, vol. 45, pp. 149–175, Jan. 1966.

[28] D. L. Snyder and I. B. Rhodes, "Some implications of the cutoff-rate criterion for coded direct detection optical communications," *IEEE Trans. on Inform. Theory*, vol. IT-26, pp. 327–338, May 1980.

[29] V. Tarokh, N. Seshadri, and A. R. Calderbank, "Space-time codes for high data rate wireless communication: performance criteria and code construction," *IEEE Trans. on Inform. Theory*, vol. IT-44, pp. 744–765, Mar. 1998.

[30] I. E. Telatar, "Capacity of multi-antenna Gaussian channels," Technical report, AT&T Bell Laboratories Technical Memo, 1995.

[31] I. E. Telatar, "Capacity of multi-antenna Gaussian channels," *European Transactions on Telecommunications*, vol. 10, no. 6, pp. 585–595, Nov.-Dec. 1999.

[32] F.-Q. Wang and D. J. Costello, "Probabilistic construction of large constraint length trellis codes for sequential decoding," *IEEE Trans. on Communications*, vol. COM-43, no. 9, pp. 2439–2448, Sept. 1995.

[33] L. Wei and H. Qi, "Near-optimal limited-search detection on ISI/CDMA channels and decoding of long convolutional codes," *IEEE Trans. on Inform. Theory*, vol. IT-46, pp. 1459–1482, July 2000.

[34] J. M. Wozencraft and R. S. Kennedy, "Modulation and demodulation for probabilistic coding," *IEEE Trans. on Inform. Theory*, vol. IT-12, pp. 291–297, July 1966.

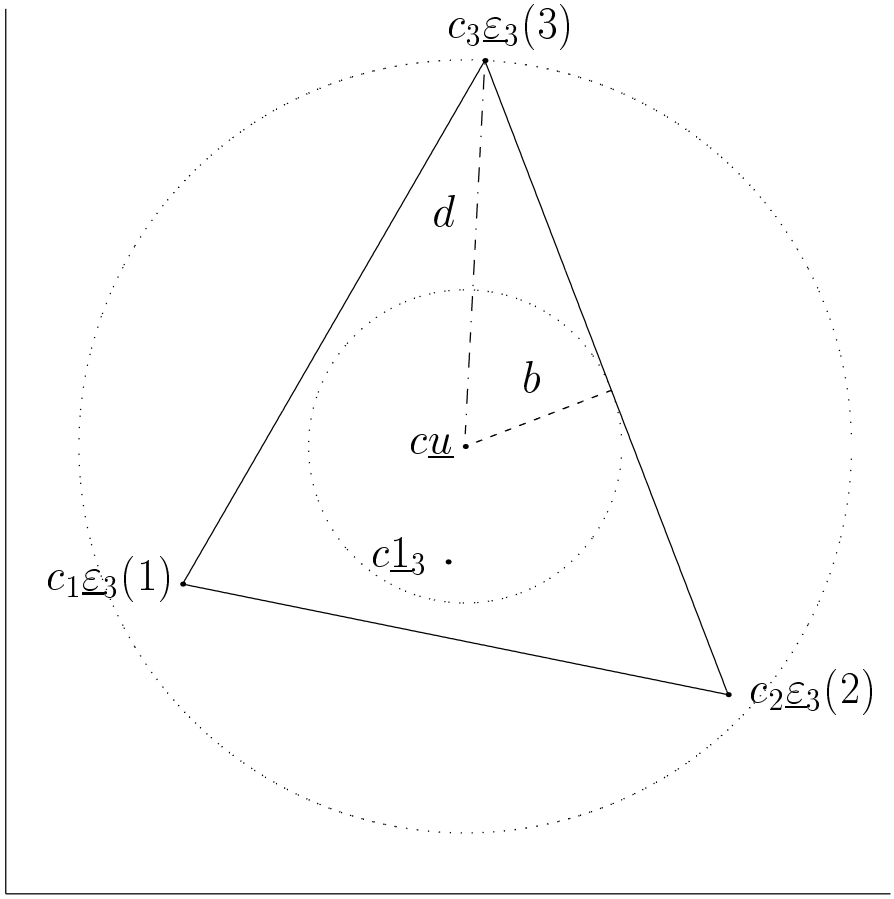

Figure 1: Two dimensional slice of three dimensional cone cone(E_k) ($k = 3$) with inscribed and circumscribed three dimensional spheres used to establish sufficient and necessary conditions, respectively, for establishing that $\underline{1}_k \in$ cone(E_k). Interior of triangle is slice of cone(E_k) in a plane orthogonal to the vector $c\underline{u}$. Vertices of triangle are located at points $\{c_i\underline{\varepsilon}_k(i)\}_{i=1}^3$ along extremal rays of cone(E_k). The inner dotted circle is the corresponding slice of the largest possible sphere centered at $c\underline{u}$ that can be inscribed in cone(E_k). The outer dotted circle is the slice of the smallest possible sphere centered at $c\underline{u}$ through which all rays in cone(E_k) must pass.

BIOGRAPHIES

Biography of Alfred O. Hero

Alfred O. Hero III, was born in Boston, MA. in 1955. He received the B.S. in Electrical Engineering (summa cum laude) from Boston University (1980) and the Ph.D from Princeton University (1984), both in Electrical Engineering. Since 1984 he has been a Professor with the University of Michigan, Ann Arbor, where he is with the Department of Electrical Engineering and Computer Science, the Department of Biomedical Engineering and the Department of Statistics. He has held visiting positions at the University of Nice at Sophia-Antipolis, France (2001), Ecole Normale Supérieure - Lyon, France (1999), Ecole Nationale Supérieure

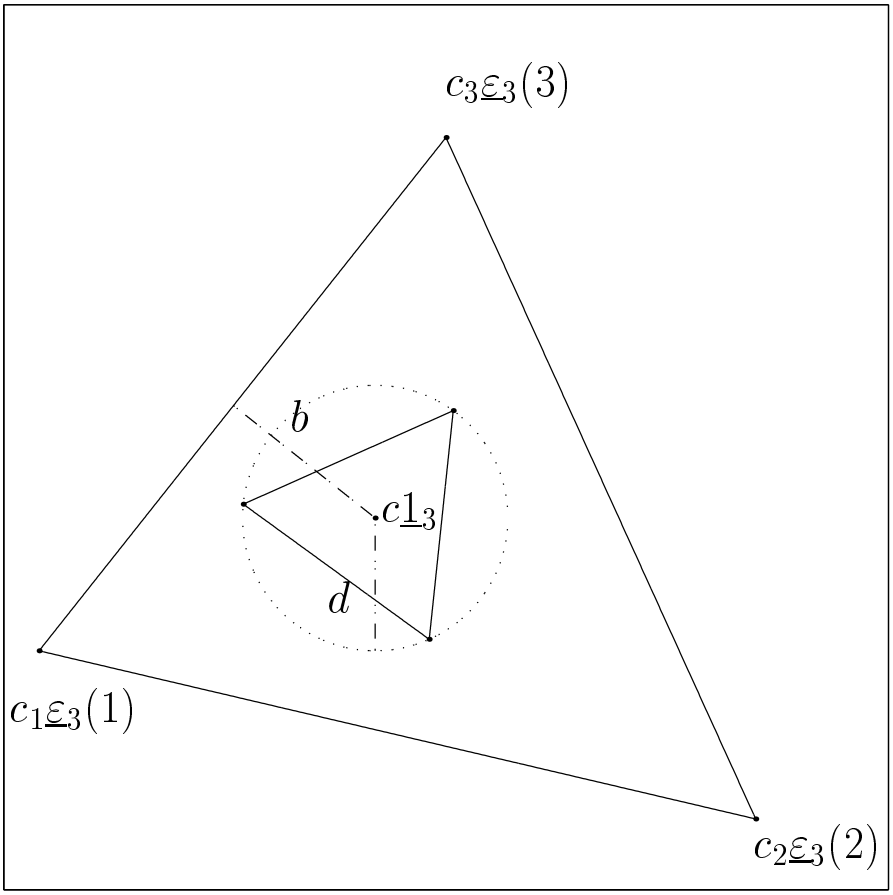

Figure 2: Interior of outer triangle is slice of cone(E_k) for $k = 3$ in a plane orthogonal to the vector $c\underline{1}_k$. Vertices of outer triangle are located at points $\{c_i\underline{\varepsilon}_3(i)\}_{i=1}^3$ along the extremal rays of cone(E_k). The interior of the smaller equilateral triangle in the center is the slice of cone$(F_k(\beta_k))$ within the same plane. The equilateral triangle has center point $c\underline{1}_k$. The dotted circle is the slice of a sphere with radius d equal to the distance between $c\underline{1}_k$ and the vertices of the smaller triangle. b is the distance of $c\underline{1}_k$ to the closest face of cone(E_k). cone$(F_k(\beta_k)) \subset$ cone(E_k) as long as the smaller triangle is inscribed in the larger triangle; i.e. $d \leq b$.

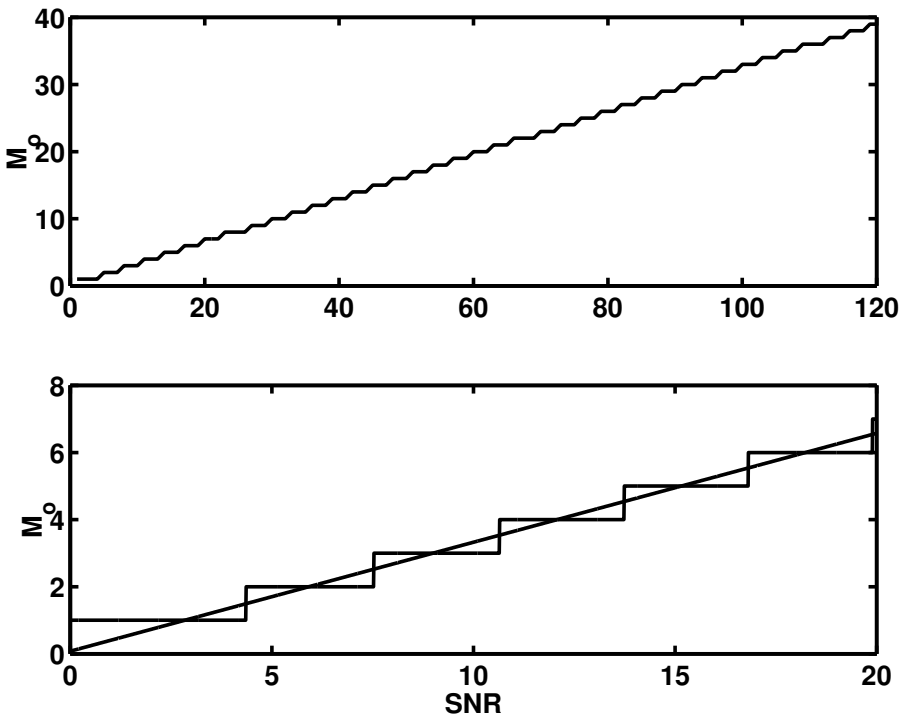

Figure 3: Top panel shows M_o given by (55) as a function of the SNR parameter ηTM. Bottom panel is blow up of first panel over a reduced range of SNR. The straight line is a least squares linear fit to the upper panel. The linear approximation has slope 0.32 and zero intercept 0.08. Average residual error between linear fit and exact M_o vs SNR step function is less than 0.09 and maximum error is less than 0.52. By Corollary 2, for T, K, M, M_o satisfying $T \geq KM_o$ and $M_o \leq M$, the curve gives the number of antennas utilized by the optimal constellation for various SNR's.

des Télécommunications, Paris, France (1999), William Clay Ford Fellow at Scientific Research Lab Ford Motor Company, Dearborn, MI (1993), Ecole Nationale des Techniques Avancees (ENSTA), Paris France (1990) and M.I.T. Lincoln Laboratory (1987 - 1989), Lexington, Massachusetts. His research interests are in the areas of estimation and detection, statistical communications, signal processing and image processing. He has served (1995-1998) as Associate Editor for the IEEE Transactions on Information Theory. He was Chairman of the Statistical Signal and Array Processing (SSAP) Technical Committee and treasurer of the Conference Board in the IEEE Signal Processing Society. He was Chairman for Publicity for the 1986 IEEE International Symposium on Information Theory (Ann Arbor, MI), General Chairman for the 1995 IEEE International Conference on Acoustics, Speech, and Signal Processing (Detroit, MI), co-chair of the 1999 IEEE Information Theory Workshop on Detection, Estimation, Classification and Filtering (Santa Fe, NM) and co-chair of the 1999 IEEE Workshop on Higher Order Statistics (Caesaria, Israel). He is currently a member of the Signal Processing Theory and Methods (SPTM) Technical Committee and Vice President (Finance) of the IEEE Signal Processing Society. He is Chair of Commission C (Signals and Systems) of the US National Commission of the International Union of Radio Science (URSI).

Alfred Hero is a Fellow of the Institute of Electrical and Electronics Engineers (IEEE), a member of Tau Beta Pi, the American Statistical Association (ASA), the Society for Industrial and Applied Mathematics (SIAM), and the US National Commission (Commission C) of the International Union of Radio Science (URSI). He has received the 1998 IEEE Signal Processing Society Meritorious Service Award, the 1998 IEEE Signal Processing Society Best Paper Award, and the IEEE Third Millenium Medal.

Biography of Thomas L. Marzetta

Thomas L. Marzetta received the Ph.D. degree in electrical engineering from the Massachusetts Institute of Technology, Cambridge.

He worked in petroleum exploration at Schlumberger-Doll Research, Ridgefield, CT, and in aerospace/electronics at Nichols Research Corporation, Wakefield, MA, before joining Bell Laboratories. He is now with the Mathematical Sciences Research Center, Bell Laboratories, Lucent Technologies, Murray Hill, NJ. The focus on his current research is high-capacity multiple-antenna wireless communications.

Dr. Marzetta has been active in the IEEE Signal Processing Society as a charter member of the Technical Committee on Multidimensional Signal Processing, as an Associate Editor of the IEEE Transactions on Acoustics, Speech, and Signal Processing, an Associate Editor of the IEEE Transactions on Image Processing,

and as a member of the Signal Processing Steering Committee of the IEEE Central New England Council. He is currently a member of the Sensor Array and Multichannel Technical Committee. He was the recipient of the 1981 ASSP Paper Award.