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Variable-Phase-Shift-Based RF-Baseband Codesign for MIMO Antenna Selection

Xinying Zhang, Andreas F. Molisch, *Fellow, IEEE*, and Sun-Yuan Kung, *Fellow, IEEE*

***Abstract*—We introduce a novel soft antenna selection approach for multiple antenna systems through a joint design of both RF (radio frequency) and baseband signal processing. When only a limited number of frequency converters are available, conventional antenna selection schemes show severe performance degradation in most fading channels. To alleviate those degradations, we propose to adopt a transformation of the signals in the RF domain that requires only simple, variable phase shifters and combiners to reduce the number of RF chains. The constrained optimum design of these shifters, adapting to the channel state, is given in analytical form, which requires no search or iterations. The resulting system shows a significant performance advantage for both correlated and uncorrelated channels. The technique works for both transmitter and receiver design, which leads to the joint transceiver antenna selection. When only a single information stream is transmitted through the channel, the new design can achieve the same SNR gain as the full-complexity system while requiring, at most, two RF chains. With multiple information streams transmitted, it is demonstrated by computer experiments that the capacity performance is close to optimum.**

I. Introduction

MULTIPLE-Input-Multiple-Output (MIMO) systems, i.e., systems that deploy multiple antenna elements at both link ends, have attracted considerable attention, due to the promise of great performance enhancements that can be achieved with them. The multiple antennas are exploited to improve the data rate and/or signal-to-noise ratio of the communication channel, as demonstrated by analytical and simulation studies [1]–[4]. However, the application of multiple antenna systems has been restricted by the increased fabrication cost and energy consumption of the RF chains (performing the microwave/baseband frequency translation) as well as the Analog-to-Digital (A/D) conversion; the number of both being linearly proportional to the number of antenna elements.

These factors motivate the recent popularity of antenna selection schemes, which "optimally" choose a subset from all the antenna elements for processing and, therefore, save the number of modulators/demodulators. With a single signal stream transmitted in the system, the hybrid selection and Maximum-Ratio Combining (MRC) approach, also known as HS-MRC, was proposed to optimize the combiner output Signal-to-Noise-Ratio (SNR). The SNR, Bit Error Rate (BER), and Symbol Error Rate (SER) performance of HS-MRC in Single-Input-Multiple-Output (SIMO) systems has been extensively investigated; see, e.g., [5]–[9]. A diversity system with selection of a *single* antenna at the transmitter and MRC at the receiver is investigated in [10]. The performance of MIMO diversity systems with antenna subset selection at *one link end* was analyzed in [11]–[13]. In [14], the joint transceiver antenna selection for a time-division multiple access (TDMA) wireless access link is considered. The algorithm and performance analysis of antenna selection combined with space-time coding have recently been addressed in [15]–[17]. When multiple signal streams are allowed for transmission in the system, antenna selection schemes and capacity bound analysis are provided in [18] and [19]; when concatenated with simple linear space-time receivers, the optimum antenna selection algorithms based on various criteria have also been derived in, e.g., [20]. The optimum branch selection often incurs a large and impractical search space. To save the computational complexity, alternative algorithms with near-optimum performance and simplified implementation are suggested by [21]–[23]. A review of the state-ofthe-art in antenna selection, including further references, can be found in [24].

Suppose we have t transmitting and r receiving antennas. In traditional approaches, the major load of signal processing is in the baseband, while the operation in the RF domain is simply an L out of r switch (for the receiver), where L is the desired number of RF links; we call this henceforth a "hard" antenna selection. Despite their great advantages in terms of cost reduction, such schemes suffer from a certain performance loss. Due to the directional nature of the multipath propagation, the signals at the antenna array are correlated. Depending on the amount of correlation, the performance of conventional selection systems can reduce to that of an L-antenna system, losing the advantage of having additional antenna elements. Even in uncorrelated channels, the performance degradation (due to smaller average SNR after link reduction) can be significant when only a small number of the antenna elements are selected. In a recent paper [25], we showed that a spatial Fourier transformation (FFT) operation in the RF domain (which can also be interpreted in the framework of the "virtual channel model" introduced in [26]), between the antenna elements and the selection switch, significantly improves the performance in strongly correlated channels. However, it shows no improvement for independently fading channels in a heavy scattering environment.

In this paper, we introduce a novel "soft" antenna selection strategy using a joint RF/baseband design. In contrast to the

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FFT scheme of [25], it uses a transformation that adapts to the channel state. More precisely, the proposed method performs an optimal r-to-L linear transformation in the RF domain, followed by L frequency down-converters and baseband processing. With soft selection, all the r receiving antennas are active, while only L demodulators are required. As the spatial diversity in all the r antennas remains fully accessible, the proposed algorithm shows significant performance improvement over the hard selection under various channel conditions. Due to practical constraints, the proposed approach uses only variable phase shifters, but no variable-gain amplifiers, in the RF domain. While cost-effective analog amplifiers in RF with satisfactory noise figure are practically not available, we stress that the rapid advances in Microwave Monolithic Integrated Circuit (MMIC) technology have made feasible the economic design and fabrication of circuitry with variable phase shifters for the microwave frequency range [27], [28].

The design of these phase shifters poses a constrained optimization problem. From a formal point of view, preprocessing for antenna selection can also be viewed as a special case of linear precoder designs [29], [30] or, more generally, precoder-decoder joint designs; see, e.g., [31]–[34]. However, the constraints arising from the limited number of downconversion chains, are quite different. We study two cases in this paper: diversity transmission and spatial multiplexing. By diversity transmission we mean the situation when all the different transmit antennas carry the weighted replicas of a single data stream. In contrast, with spatial multiplexing, different data streams can be applied on the t transmitting antenna elements to provide a maximal data rate. We will prove that for diversity transmission, the new scheme can achieve the full SNR gain in the channel as long as more than two branches are used $(L \geq 2)$. When $L = 1$, the SNR gain of our new scheme is also well above the conventional selection. The proposed scheme is robust for various channel conditions and can be applied at both the receiver and the transmitter ends. A similar structure is proposed for the case of spatial multiplexing. The theoretical analysis shows that to maximize the system capacity for a given number of transceiver chains, a symmetric selection should be adopted, i.e., an L-out-of-t selection in the transmitter and an L-out-of-r selection in the receiver side; the system can support at most L independent information streams in this case. The unconstrained optimum transmitter selection is the power allocation of L-fold water-filling in the baseband, followed by a transformation to the eigenspace of the channel in the RF domain. A closed-form phase-shift-based approximation is given for the RF operations as a suboptimum solution to the constrained optimization of capacity. Computer experiments demonstrate that the proposed scheme delivers a capacity very close to the optimum choice.

A. Organization

The rest of the paper is organized as follows. The theoretical analysis of the antenna selection strategy and optimum design of the phase shifters are addressed in Sections II and III for diversity transmission and spatial multiplexing, respectively. In these two sections, the antenna selection design is conducted based on a fixed channel realization, and there is no constraint on the code length. The stochastic performance (in terms of, e.g., average SNR gain and outage capacity) of the algorithm in practical environments is then investigated through analysis and simulations in Section IV based on the stochastic channel model introduced therein.

B. Acronyms

MRT	Maximum-Ratio Transmission.
MRC	Maximum-Ratio Combining.
FC	Full-Complexity.
HS	Hybrid Selection.
FFTS	FFT-based Selection.
PSS	Phase-Shift and Selection.
RLC	Receiver Linear Combining.
TLC	Transmitter Linear Combining.
TRLC	Transmitter-and-Receiver Linear Combining.

II. Antenna Selection Under Diversity Transmission

A MIMO system model for diversity transmission is depicted in Fig. 1(a). H is the $r \times t$ matrix denoting the transfer function of the MIMO channel. The original source stream $s(k) \in \mathcal{C}$ is multiplied by the t-dimensional complex weighting vector $\vec{v}$, before it is modulated to the passband and applied to each of the t transmitting antennas. The sampling vector $\vec{x}(k) \in \mathcal{C}^r$ of the receiver observations is

$$\vec{x}(k) = H\vec{v}s(k) + \vec{n}(k). \tag{1}$$

The total transmission power is constrained to P: $\mathcal{E}[s(k)s^*(k)] \leq P$. The thermal noises $\vec{n}(k) \in \mathcal{C}^r$ are zero-mean, i.i.d. circularly symmetric Gaussian random processes with variance $\sigma_n^2 I_r$: $\vec{n}(k) \sim \mathcal{N}_C(\vec{0}, \sigma_n^2 I_r)$. To facilitate the performance evaluation, we define the nominal SNR to be $\rho = P/\sigma_n^2$. The t-dimensional transmitter weighting vector satisfies $\|\vec{v}\| = 1$.

With diversity transmission, either all the r receiver streams (with full-complexity reception) or a subset of these signals (with antenna selection) are demodulated and exploited for the estimation of the source information. The SNR gain, which is defined as the ratio of combiner output SNR at the receiver to the nominal receiver SNR $\rho\mathcal{E}[|H_{ij}|^2]$, serves as the major measure for the performance. We assume that perfect channel state information (CSI) is available at the receiver as well as the transmitter. For simplicity, the antenna selection at receiver side will be introduced in Section II-B. The transmitter selection can be designed in a similar fashion by observing the duality between transmitter and receiver processing. The joint transceiver design for antenna selection will be addressed in Section II-C.

A. Full-Complexity Maximum-Ratio Combining (FC-MRC) Reception

The full-complexity reception consists of two blocks [see Fig. 1(a)]: r demodulators and RF chains and an r-to-1 RLC (Receiver Linear Combining) block in baseband. The RLC block outputs the signal estimate $\hat{s}(k)$ resultant from a weighting vector $\vec{u}$:

$$\hat{s}(k) = \vec{u}^* H\vec{v}s(k) + \vec{u}^*\vec{n}(k) \tag{2}$$

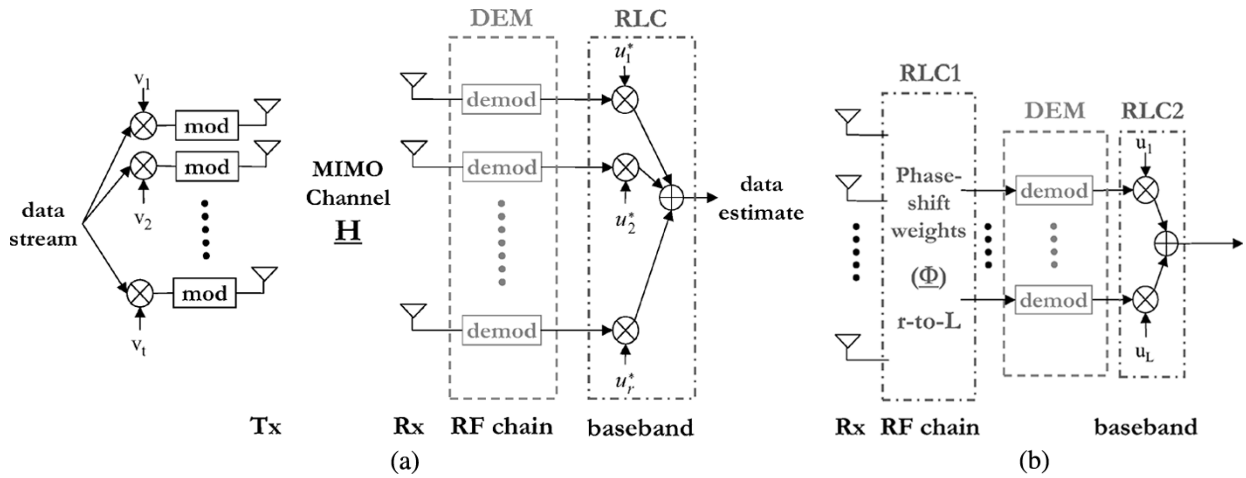


Fig. 1. (a) MIMO system diagram with diversity transmission and FC-MRC reception. (b) PSS-MRC reception with an r-to-L ($L < r$) RLC with phase-shift-only weights in RF chains and L demodulators.

where superscript $*$ denotes the conjugate transpose of a matrix or vector (for a scalar, it reduces to the complex conjugation). The average SNR of $\hat{s}(k)$ in the equation above is

$$\frac{\mathcal{E}\left[|\vec{u}^* H\vec{v}s(k)|^2\right]}{\mathcal{E}\left[|\vec{u}^*\vec{n}(k)|^2\right]} = \rho\frac{|\vec{u}^* H\vec{v}|^2}{\|\vec{u}^*\|^2}. \tag{3}$$

For the determination of the optimum weights, we introduce the singular value decomposition (SVD) of H: $H = U_H\Sigma_H V_H^*$, where U_H and V_H are $r\times r$, $t\times t$ unitary matrices representing the left and right singular vector spaces of H, respectively, and $\Sigma_H \in \mathcal{R}^{r\times t}$ is nonnegative and diagonal, consisting of all the singular values of H. For convenience, we denote $\lambda_{H,i}$ as the ith largest singular value of matrix H, and $\vec{u}_{H,i}(\vec{v}_{H,i})$ is the left (right) singular vector of H associated with $\lambda_{H,i}$. To maximize the estimate SNR term in (3), the optimum weights are known as MRT and MRC [13], [35], [36], adapting to the channel coefficients: $\vec{u} = \vec{u}_{H,1}, \vec{v} = \vec{v}_{H,1}$. The resultant optimal SNR gain is therefore [13], [35], [36]

$$\gamma_{FC} = \frac{\left|\vec{u}_{H,1}^* H\vec{v}_{H,1}\right|^2}{\left\|\vec{u}_{H,1}^*\right\|^2} = \lambda_{H,1}^2. \tag{4}$$

It is well known that FC-MRC delivers the full diversity order of rt in addition to SNR gain.

B. Receiver Design: PSS-MRC (Phase-Shift and Selection MRC) Reception

The efficient use of onboard energy is demanded for a real hardware implementation. Intuitively, one could minimize the demodulation cost in FC-MRC reception by switching the operational order of demodulators and RLC. As the RLC block here functions as an r-to-1 switch in RF, only one demodulator is required afterwards. However, in practice, many existing antenna systems may not possess the capability for cost-effective amplitude adjustments with low power consumption [37]. On the other hand, the rapid advances in MMIC techniques enable the economic design and fabrication of pure variable phase shifters for the microwave frequency range [27], [28]. It is thus desirable to use only variable phase-shifters in the RF domain to simplify hardware design. This leads to a constrained optimization problem, with $|v_i| = 1$.

Along this line, our proposed Phase-Shift and Selection MRC (PSS-MRC) scheme adopts the architecture in Fig. 1(b): The load of the optimal MRC combining $\vec{u} = \vec{u}_{H,1}$ is divided into two linear steps, namely, RLC1 and RLC2, in RF and baseband, respectively. Only pure phase-shifters and adders are allowed in RCL1, which serves as an r-to-L ($L < r$) switch with L output streams for demodulation. RLC2 is an L-to-1 linear combiner with any desirable weights to produce the signal estimate. The major demodulation complexity in the new system is now proportional to L; on the other hand, a smaller L could also possibly result in more information loss in the r-to-L switch RLC1. The tradeoff between hardware complexity and SNR performance of the PSS-MRC system is addressed below.

The r-to-L RLC1 switch in Fig. 1(b) is equivalent to an $L\times r$ matrix Φ multiplied on the data vectors $\vec{x}(k)$. All the elements in Φ are nonzero and restricted to be pure phase-shifters. To investigate the optimal design of the phase-shift matrix Φ and the resultant SNR gain, we denote the set of all such matrices as $\mathcal{F}^{L\times r} = \{\Phi | [\Phi]_{m,n} = e^{j\phi_{m,n}}; 1 \le m \le L, 1 \le n \le r, \phi_{m,n} \in \mathcal{R}\}$. The RLC2 operation is an L-dimensional vector $\vec{u} \in \mathcal{C}^{L\times 1}$. The cascade of RLC1, RLC2 is again a linear combiner $\vec{u}^*\Phi$, and the optimal post SNR gain at the output of RLC2 is

$$\gamma_{PSS}(L) = \max_{\Phi\in\mathcal{F}^{L\times r}} \max_{\vec{u}\in\mathcal{C}^{L\times 1}} \max_{\|\vec{v}\|=1} \frac{|\vec{u}^*\Phi H\vec{v}|^2}{\|\vec{u}^*\Phi\|^2}. \tag{5}$$

Comparing (3) with (5), it can be deduced that the PSS-MRC can achieve the same SNR gain as FC-MRC if and only if there exists a solution of Φ and $\vec{u}$ to the following constrained problem:[1]

$$\vec{u}^*\Phi = \vec{u}_{H,1}^*; \quad \Phi \in \mathcal{F}^{L\times r}, \quad \vec{u} \in \mathcal{C}^{L\times 1}. \tag{6}$$

Theorem 1 (Optimal SNR Gain of PSS-MRC): For a general $r\times t$ MIMO channel H, there exists a solution to (6) if and only if $L \ge 2$. In other words, the PSS-MRC scheme could deliver the same SNR gain as FC-MRC: $\gamma_{PSS}(L) = \gamma_{FC} = \lambda_{H,1}^2$ if and only if no less than two demodulators are allowed: $L \ge 2$. The optimum weights of $\Phi \in \mathcal{F}^{L\times r}$, $\vec{u} \in \mathcal{C}^L$ for achieving this gain depend on the channel H.

Proof: See Appendix A. ■

[1]Here it is assumed that the same MRT weights as in the FC-MRC scheme is adopted in the transmitter side.

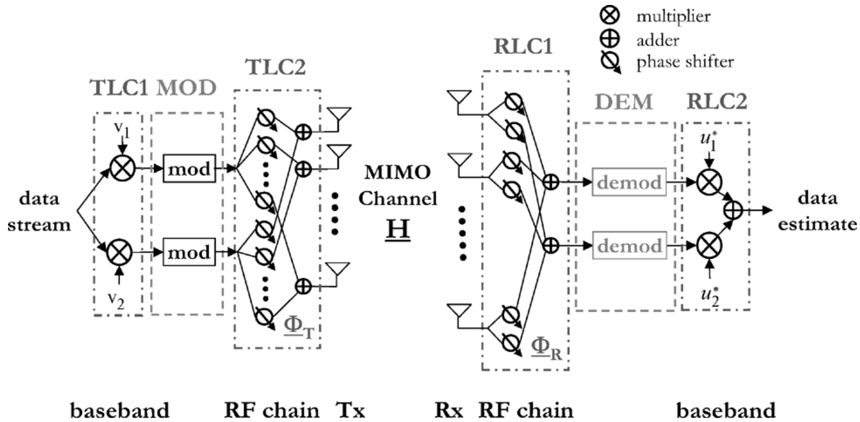


Fig. 2. System diagram of the transmitter-receiver co-design for PSS-MRC/MRT scheme, with $L_T = L_R = 2$.

The theorem above shows that involving more than two demodulators does not help to improve the SNR gain. To complete the analysis, we continue with the case when only $L = 1$ demodulator is available in the system. For $L = 1$, the optimization problem in (5) is constrained to the so-called r-torus $\mathcal{F}^{1\times r}$ [37], and the formulation is also known as equal gain combining (EGC) [38]. The optimum solution can be obtained by taking partial derivatives to form a set of r nonlinear differential equations, and the globally optimal point is generally not available in closed-form; in fact, the SNR gain has many local maxima in the torus. Alternatively, a lot of iterative algorithms in the literature of analog beam forming and steering can be applied; a quantized EGC algorithm is also proposed in [38], which quantizes the combining vector space with finite elements and finds the optimal solution via the brute-force search in the quantized space. All of these algorithms require an exhaustive search step; therefore, to minimize the computational complexity, we provide a suboptimal solution in closed form for the PSS design when $L = 1$. Recalling (5), the optimum SNR gain is now

$$\begin{aligned}\max_{\Phi\in\mathcal{F}^{1\times r}} \max_{u\in\mathcal{C}} \max_{\|\vec{v}\|=1} & \frac{|u^*\Phi U_H\Sigma_H V_H^*\vec{v}|^2}{\|u^*\Phi\|^2} \\ &= \max_{\Phi\in\mathcal{F}^{1\times r}} \max_{\|\vec{v}\|=1} \frac{|\Phi U_H\Sigma_H V_H^*\vec{v}|^2}{r} \\ &= \max_{\Phi\in\mathcal{F}^{1\times r}} \frac{|\Phi U_H\Sigma_H V_H^* V_H\Sigma_H^* U_H^*\Phi^*|^2}{r\,\|V_H\Sigma_H^* U_H^*\Phi^*\|^2} \\ &= \max_{\Phi\in\mathcal{F}^{1\times r}} \frac{1}{r}\sum_{i=1}^{k_H} |\Phi\vec{u}_{H,i}|^2\lambda_{H,i}^2 \qquad (7)\end{aligned}$$

where k_H is the rank of H: $k_H \le \min(t, r)$.

As $\lambda_{H,1} \ge \lambda_{H,2} \ge \ldots \ge \lambda_{H,k_H}$, the maximization of (7) can empirically be approximated by the optimization of $|\Phi\vec{u}_{H,1}|^2$. We denote the ith element of $\vec{u}_{H,1}$ as $\beta_{i,1}e^{j\varphi_{i,1}}$, then

$$\begin{aligned}\max_{\Phi\in\mathcal{F}^{1\times r}} |\Phi\vec{u}_{H,1}|^2 &= \max_{\Phi\in\mathcal{F}^{1\times r}} \left|\sum_{i=1}^{r}\beta_{i,1}e^{j\phi_i}e^{j\varphi_{i,1}}\right|^2 \\ &= \left(\sum_{i=1}^{r}\beta_{i,1}\right)^2 \quad \text{with } \Phi = [e^{-j\varphi_{1,1}}\ldots e^{-j\varphi_{r,1}}].\end{aligned}$$

The phase-shifters in the PSS-MRC approach are designed by extracting and reversing the phases of $\vec{u}_{H,1}$. The advantage of the proposed PSS-MRC scheme will be further confirmed by the analysis and simulations in Section IV. It will be shown that, with high spatial correlation of channel fadings, such a design is optimum and delivers the same SNR gain as the FC-MRC reception; in a more general MIMO environment, the SNR performance is also near optimum.

C. Transceiver Co-Design: PSS-MRC/MRT Scheme

The proposed PSS-MRC reception is focused on the receiver side, with the underlying assumption that FC-MRT is adopted in the transmitter side. To further reduce the hardware complexity, a total system solution requires the joint design of both the transmitter and receiver.

In the full-complexity design, the optimum MRT and MRC weighting vectors share a similar pattern: $\vec{u} = \vec{u}_{H,1}$ and $\vec{v} = \vec{v}_{H,1}$. Therefore, the two ends can be treated in duality and the same PSS strategy can be applied on the transmit antennas. The system diagram of the joint PSS design is exemplified in Fig. 2 with $L_T = 2$ modulators and $L_R = 2$ demodulators. Similar to the PSS reception, the transmitter side also consists of two linear combiners: TLC1 is a 1-to-L_T $(L_T < t)$ switch in baseband, which divides the original information stream into L_T branches with the weights $\vec{v}$; TLC2 is an L_T-to-t microwave switch with only phase shifters and adders. The receiver structure is the same as before. We denote the best SNR gain of such a system as $\gamma_{PSS}(L_T, L_R)$. When $L_T \ge 2$, $L_R \ge 2$, following the argument in the proof of Theorem 1, we can always achieve the MRT/MRC weights by the cascade of the linear combiners at the two ends: $\Phi_T\vec{v} = \vec{v}_{H,1}$ and $\vec{u}^*\Phi_R = \vec{u}_{H,1}^*$. The phases of Φ_T can be determined from (35) by replacing $(\beta_{i,1}, \varphi_{i,1})$ with the magnitudes and phases of the elements in $\vec{v}_{H,1}$. The post SNR gain satisfies

$$\gamma_{PSS}(L_T, L_R) = \gamma_{FC}, \quad L_T \ge 2, \quad L_R \ge 2. \qquad (8)$$

When $L_T = 1$, i.e., only one modulator in the transmitter, the weights are chosen similarly as before: the phases of Φ_T are the negative of the phases in $\vec{v}_{H,1}$; as the transmitter weights must satisfy the power constraint $\|\Phi_T v\| = 1$, the weight in RLC1 is selected as $v = 1/\sqrt{t}$.

D. Unification of Various Selection Schemes

Below, we give a brief overview of some existing antenna selection schemes with comparisons in a unified subspace framework. For simplicity, only receiver antenna selection is considered, assuming FC-MRT at the transmitter.

1) Hybrid Selection MRC (HS-MRC): In conventional antenna selection, L out of the r receive antennas are chosen for communication. Mathematically, each selection option corresponds to an $L \times r$ selection matrix S on the transfer function, which extracts L rows from H that are associated with the selected antennas. We denote $\mathcal{S}_L$ as the set of all such selection matrices. For any selection option, the optimal estimate SNR is achieved via a similar MRC on the L branches:

$$\max_{\vec{u}\in\mathcal{C}^L}\max_{\|\vec{v}\|=1}\frac{\|\vec{u}^*SH\vec{v}\|^2}{\|\vec{u}\|^2}=\lambda^2_{SH,1}. \tag{9}$$

The HS-MRC chooses the optimal selection matrix S that maximizes the estimate SNR above:

$$\gamma_{\mathrm{HS}}(L)=\max_{S\in\mathcal{S}_L}\lambda^2_{SH,1}. \tag{10}$$

2) FFT-based Selection (FFTS): To cope with the highly correlated MIMO channels, a FFT-based selection scheme is proposed in [25], where an r-point FFT matrix Φ_{FFT} is inserted in the RF chains before the selection S (Φ_{FFT} is normalized: $\Phi_{\mathrm{FFT}}\Phi^*_{\mathrm{FFT}}=I_r$ to preserve the noise level). Via a similar MRC on the L selected branches after FFT, the optimal estimate SNR in FFTS is:

$$\begin{aligned}\gamma_{\mathrm{FFTS}}(L)&=\max_{S\in\mathcal{S}_L}\max_{\vec{u}\in\mathcal{C}^L}\max_{\|\vec{v}\|=1}\frac{|\vec{u}^*S\Phi_{\mathrm{FFT}}H\vec{v}|^2}{\|\vec{u}\|^2}\\&=\max_{S\in\mathcal{S}_L}\lambda^2_{S\Phi_{\mathrm{FFT}}H,1}.\end{aligned} \tag{11}$$

The aforementioned four schemes (including FC-MRC, HS-MRC, FFTS and PSS) can be unified from a subspace perspective. The optimal SNR gains in (4), (5), (10), and (11) can be equivalently written as follows:

$$\max_{\vec{u}\in\mathcal{U}}\max_{\|\vec{v}\|=1}\rho\frac{|\vec{u}^*H\vec{v}|^2}{\|\vec{u}\|^2} \tag{12}$$

with different constraint sets $\mathcal{U}$:

1) FC-MRC: $\mathcal{U}_{FC}=\mathcal{C}^{r\times 1}$;
2) HS-MRC: $\mathcal{U}_{\mathrm{HS}}(L)=\cup_{S\in\mathcal{S}_L}R.S.(S)$, where $R.S.(S)$ is the row span of S;
3) FFTS: $\mathcal{U}_{\mathrm{FFTS}}(L)=\cup_{S\in\mathcal{S}_L}R.S.(S\Phi_{\mathrm{FFT}})$;
4) PSS: $\mathcal{U}_{PSS}(L)=\cup_{\Phi\in\mathcal{F}^{L\times r}}R.S.(\Phi)$.

The four candidate spaces are related by $\mathcal{U}_{\mathrm{HS}}(L),\mathcal{U}_{\mathrm{FFTS}}(L)\subset\mathcal{U}_{PSS}(L)\subset\mathcal{U}_{FC}$. Therefore, the estimate SNR gains are ordered as follows:

$$\gamma_{\mathrm{HS}}(L),\gamma_{\mathrm{FFTS}}(L)\le\gamma_{PSS}(L)\le\gamma_{FC}. \tag{13}$$

Due to the significant space expansion from $\mathcal{U}_{\mathrm{HS}}(L)$ to $\mathcal{U}_{FC}$, FC-MRC performs much better than HS-MRC under most channel conditions: $\gamma_{FC}\gg\gamma_{\mathrm{HS}}(L)$. In some practical MIMO channels, the performance of HS-MRC can reduce to that of an L-antenna system, losing the advantage of having additional antenna elements.

The advantage of FFTS is significant in strongly correlated channels (e.g., small angular spread, small antenna spacing, LoS propagation). In this case, the signal energy arriving at the receiver array is concentrated to a small angle, and the output of the FFT matrix transforms the receiver signals into the angular space, in which the energy is only focused on very few number of beams regarding the directions in space. Therefore, after selection, FFTS should have the near optimum SNR gain: $\gamma_{\mathrm{HS}}(L)\ll\gamma_{\mathrm{FFTS}}(L)\approx\gamma_{FC}$. A channel with weak correlations may have large angular spread, large antenna spacing, or heavy scattering in the environment. The fading coefficients are almost independent of each other, and the angular spectrum has a wide spread. In this case, FFT has little effect on the performance as the Power Azimuth Spectrum (PAS) remains approximately uniform after FFT, namely, $\gamma_{\mathrm{HS}}(L)\approx\gamma_{\mathrm{FFTS}}(L)\ll\gamma_{FC}$.

As for PSS-MRC, recalling the results in Theorem 1, we have $\mathcal{U}_{PSS}(L)=\mathcal{U}_{FC}$ when $L\ge 2$, which further implies that $\gamma_{PSS}(L)=\gamma_{FC}$ for $L\ge 2$. Simulation results will be provided in Section IV to support the analysis. It is shown that PSS scheme shows a significant improvement over both HS and FFTS. It delivers a satisfactory SNR gain under both strongly correlated and independent channel conditions.

III. Antenna Selection for Spatial Multiplexing Transmission

In this section, we study the case of spatial multiplexing. The analysis and results in this section are related to the work in the precoding literature [29], [32], [39] but refer more explicitly to the use of antenna selection. With spatial multiplexing, the tranceiver system model is

$$\vec{x}(k)=H\vec{s}(k)+\vec{n}(k) \tag{14}$$

where $\vec{s}(k)$ is now a $t\times 1$ vector denoting the different transmit sequences. Capacity is a major performance measurement for spatial multiplexing system, which indicates the maximal information rate supported by the system with error-free transmission. Two scenarios will be separately considered: 1) The channel state information (CSI) is unknown to the transmitter, and 2) the CSI is available at the transmitters. Without channel information, the antenna selection is mainly focused on the receiver side as described in Section III-A; in the latter case with CSI available, a joint selection including processing at both ends is desired, which leads to the optimum transceiver co-design in Section III-B. In this section, the channel is assumed to be flat-fading and quasistatic, i.e., the coherence time of the channel is so long that a large number of bits can be transmitted within this time. More specifically, we assume that the data are encoded with near Shannon limit achieving codes. Thus, each channel realization can be associated with a (Shannon—AWGN) capacity value. The capacity thus becomes a random variable, rendering the concept of a *capacity cumulative distribution function (cdf)* a meaningful performance measure.

A. Antenna Selection Without Transmitter Design

When the channel fading coefficients are unavailable at the transmitter, we assume a uniform power distribution at the t transmit antennas: $\mathcal{E}[\vec{s}(k)\vec{s}^*(k)] = (P/t)I_t$. Without antenna selection, the full-complexity system with r demodulators delivers a capacity of

$$C_{FC} = \log_2 \det \left(I_t + \frac{\rho}{t} H^* H\right) = \sum_{i=1}^{k_H} \log_2 \left(1 + \frac{\rho}{t}\lambda_{H,i}^2\right) \quad (15)$$

when assuming i.i.d. Gaussian input streams.

Similar to the PSS-MRC design for diversity transmission in Section II, we propose a soft antenna selection structure by inserting a linear r-to-L transformer in the RF domain. We will start with the unconstrained optimal antenna selection problem without restriction on the weighting matrix $R \in \mathcal{C}^{L\times r}$; a phase-shift-based antenna selection method constrained to $R \in \mathcal{F}^{L\times r}$ will be introduced afterwards to reduce the RF chain processing cost.

For a MIMO system with the linear transformation $R \in \mathcal{C}^{L\times r}$ in receiver RF links, the mutual information between the two ends of the system (with colored noise) is [40]

$$I(\vec{s}, \vec{x}) = \log_2 \det \left[I_t + \frac{\rho}{t} H^* R^* (RR^*)^{-1} RH\right]. \quad (16)$$

The capacity is the maximization of the mutual information over all valid R:

$$C_{RLC}(L) = \max_{R\in\mathcal{C}^{L\times r}} \log_2 \det \left[I_t + \frac{\rho}{t} H^* R^* (RR^*)^{-1} RH\right]. \quad (17)$$

As can be shown along lines that are similar to [1], the optimum RLC weights that maximize the capacity above is summarized in the following theorem.

Theorem 2 (Optimum r-to-L RLC Antenna Selection): The solution to the unconstrained optimization problem in (17) is

$$C_{RLC}(L) = \sum_{i=1}^{L} \log_2 \left(1 + \frac{\rho}{t}\lambda_{H,i}^2\right) \quad (18)$$

with the optimum choice

$$R = B[\vec{u}_{H,1} \quad \dots \quad \vec{u}_{H,L}]^* \quad (19)$$

where $\lambda_{H,i}$, $\vec{u}_{H,i}$ are the singular values and vectors of H as defined before, and B is any $L \times L$ nonsingular matrix.

Proof: See Appendix B ■

Comparing (15) and (18), if the desired number of demodulators $L < k_H$ (which means, necessarily, $L < t$ and $L < r$), we always have $C_{RLC}(L) < C_{FC}$. On the other hand, if L is selected such that $L \geq k_H$, the RLC selection can always achieve the same capacity as the full-complexity scheme. The optimal RLC matrix R is to project the observations into the eigen-space of HH^* associated with the L largest singular values. It is also implied by (18) that to achieve the capacity of the full-complexity MIMO system, the sufficient and necessary number of demodulators required for the RLC selection system is $L = k_H \leq \min(t, r)$.

The unconstrained optimization result in Theorem 2 facilitates the phase-shifter design of the PSS system, with the RF weighting matrix consisting of phase-shift-only elements. The best choice of PSS is formulated as the following constrained optimization problem:

$$\max_{\Phi\in\mathcal{F}^{L\times r}} \log_2 \det \left[I_t + \frac{\rho}{t} H^* \Phi^* (\Phi\Phi^*)^{-1} \Phi H\right]. \quad (20)$$

Again, as a closed-form solution is not available, the phase-shifters of Φ are chosen in the same manner as in the diversity transmission part with $L = 1$. Namely, the phases of the unconstrained optimum RLC matrix R in (19) are extracted to form the weight matrix Φ: denoting the kth element of vector $\vec{u}_i(H)$ as $\beta_{k,i}e^{j\varphi_{k,i}}$, we choose the (k,i)th element $e^{j\phi_{k,i}}$ in matrix Φ as $\phi_{k,i} = -\varphi_{k,i}$.

B. Antenna Selection With Transceiver Co-Design

When the CSI is available at both ends, with all the r branches, the capacity of the original spatial multiplexing channel in (14) is

$$C_{FC} = \max_{p(\vec{s}(k))} I(\vec{s}(k); \vec{x}(k)) = \max_{p(\vec{s}(k))} \log_2 \det \left\{I_r + \frac{1}{N} H \mathcal{E}\left[\vec{s}(k)\vec{s}^*(k)\right] H^*\right\} \quad (21)$$

where $p(\vec{s}(k))$ is the probability density function (pdf) of $\vec{s}(k)$. The optimal choice of $\mathcal{E}[\vec{s}(k)\vec{s}^*(k)]$ is known as water-filling [1], when the different transmitting signals are complex Gaussian sequences with the variance $\mathcal{E}[\vec{s}(k)\vec{s}^*(k)] = V_H \, \mathrm{diag}_{i=1}^t\{(P/t)[\mu - (t/\rho\lambda_i^2(H))]^+\} \, V_H^*$. Here, μ is a constant satisfying the power constraint $\sum_{i=1}^{k_H} [\mu - (t/\rho\lambda_{H,i}^2)]^+ = t$, and $[a]^+$ is defined as $\max(a, 0)$.

With CSI available at both ends, the antenna selection strategy can be jointly applied at the transmitter and the receiver through a transceiver co-design. We first address the case when there is no constraint on the linear RF transformations, and the phase-shift-based approach will follow momentarily. The **unconstrained** joint antenna selection design is formulated as

$$C_{\mathrm{TRLC}}(L) = \max_{R\in\mathcal{C}^{L\times r}} \max_{\mathrm{tr}[TT^*]\leq t} \log_2 \det\left[I_t + \frac{\rho}{t}(RHT)^*(RR^*)^{-1}RHT\right] \quad (22)$$

where T and R denote the transmitter and receiver selection matrices that form a Transmitter-and-Receiver Linear Combining (TRLC) pair. It should be noted that in the formulation above, we fix the number of RF links after selection to be L at the receiver side, while the number of RF links at the transmitter is not fixed. Namely, T can be any $t \times m$ selection matrix, where $m \leq t$. The subsequent analysis shows that with the L-out-of-r selection in the receiver side, for the optimal capacity performance, at most L independent streams can be transmitted through the system.

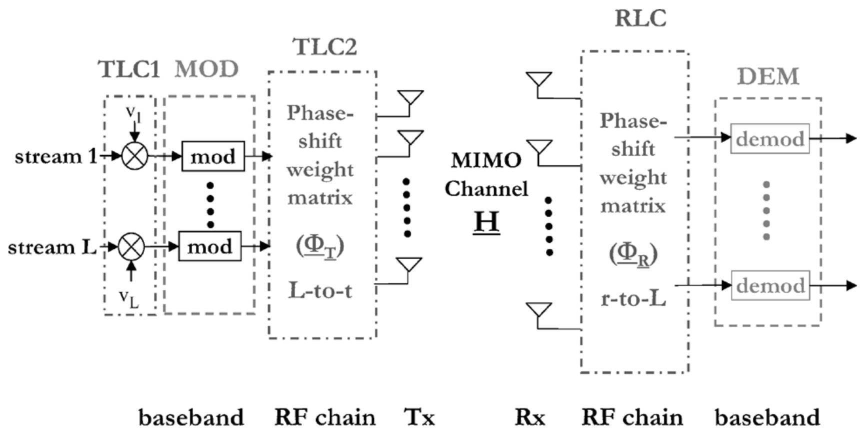


Fig. 3. Phase-Shift and selection with transmitter and receiver co-design under spatial multiplexing transmission.

Theorem 3 (Optimal Joint TRLC Antenna Selection): The solution to the **unconstrained** optimization problem in (22) is

$$C_{\mathrm{TRLC}}(L) = \sum_{i=1}^{L} \left[\log_2\left(\frac{\rho}{t}\mu\lambda_{H,i}^2\right)\right]^+. \tag{23}$$

One choice of T and R that can achieve the capacity above is

$$T = [\vec{v}_{H,1} \quad \dots \quad \vec{v}_{H,L}]\mathrm{diag}_{i=1}^{L}\left[\sqrt{\left(\mu - \frac{t}{\rho\lambda_{H,i}^2}\right)^+}\right] A$$
$$R = B[\vec{u}_{H,1} \quad \dots \quad \vec{u}_{H,L}]^*. \tag{24}$$

Here, A can be any $L \times L$ unitary matrix, B can be any full-rank matrix of the same size, and the constant μ satisfies $\sum_{i=1}^{L}(\mu - (t/\rho\lambda_{H,i}^2))^+ = t$.

Proof: See Appendix C

Based on the **unconstrained** optimum selection matrix above, a joint transceiver design with **phase-shift constraint** in the RF domain is proposed to approximate the optimum transformations. The total system structure is plotted in Fig. 3, which is referred to as the PSS scheme. The transmitter consists of L modulators and two TLC blocks. The first TLC block before modulation is to allocate the water-filling power on the L independent streams, according to the $L \times L$ diagonal factor of T in (24). The second TLC is an L-to-t linear combiner. The Lt variable phase shifters of Φ_T are extracted from the optimum TRLC weight matrix $T = [\vec{v}_{H,1} \ \dots \ \vec{v}_{H,L}]$. To meet the power constraint $\mathrm{tr}[\mathrm{diag}(\vec{v})^*\Phi_T^*\Phi_T(\mathrm{diag}(\vec{v})] \le t$, the power allocation weights $\vec{v}$ in TLC1 are proportionally modified by a constant $(1/\sqrt{t})$: $v_i = (1/\sqrt{t})\sqrt{(\mu - (t/\rho\lambda_{H,i}^2))^+}$. The receiver side follows the same design rule as presented in Section III-A.

IV. Simulation

The theoretical findings are supported by the performance analysis and simulation results presented in this section. Various Monte Carlo tests are conducted for the two scenarios: 1) diversity transmission and 2) spatial multiplexing transmission. The performances of different antenna selection techniques, including Full-Complexity (FC), Hybrid Selection (HS), FFT-based Selection (FFTS), and Phase-Shift and Selection (PSS) schemes, are compared. We adopt the spatially correlated channel model that has been extensively used in [16], [41]:

$$H = C_R^{\frac{1}{2}} W C_T^{\frac{1}{2}} \tag{25}$$

where W is a Rayleigh fading matrix with i.i.d. circularly symmetric complex Gaussian entries $\sim \mathcal{N}_{\mathcal{C}}(0,1)$, and C_R, C_T are $r \times r$, $t \times t$ matrices denoting receive and transmit correlations, respectively. The correlation matrices C_T and C_R are determined by the AoA (Angle-of-Arrival) and AoD (Angle-of-Departure). The Gaussian distributed PAS of the AoA follows $\theta = \theta_R + \epsilon$; $\epsilon \sim \mathcal{N}(0, \sigma_R^2)$. When the angle spread σ_R is small, and with a ULA (Uniformly-spaced Linear Antenna) array at the receiver side, the assumptions above allow a closed-form computation of C_R and C_T, as given in [42].

A. SNR Performance Under Diversity Transmission

We first investigate the performance of antenna selection in the strongly correlated channel, in which the PAS consists of one explicit spatial direction with no angle spread ($\sigma_T = \sigma_R = 0$). This simplification provides a close approximate to more general channels where the angle spreads are small. The bounds of γ_{FC} for Rayleigh fading MIMO channels can be found in some earlier literatures, e.g., [43].

When $\sigma_T = \sigma_R = 0$, the correlation matrices C_T, C_R collapse to two rank-1 matrices:

$$C_T = \vec{a}_T(\theta_T)\vec{a}_T^*(\theta_T), \quad C_R = \vec{a}_R(\theta_R)\vec{a}_R^*(\theta_R)$$

where $\vec{a}_T(\theta)$ and $\vec{a}_R(\theta)$ are the transmit/receive antenna response vectors defined as

$$\vec{a}_T(\theta) = \begin{bmatrix}1 & e^{j2\pi d_T \cos\theta} & \dots & e^{j2\pi(t-1)d_T\cos\theta}\end{bmatrix}^T$$
$$\vec{a}_R(\theta) = \begin{bmatrix}1 & e^{j2\pi d_R \cos\theta} & \dots & e^{j2\pi(r-1)d_R\cos\theta}\end{bmatrix}^T$$

and d_T (d_R) is the relative receive (transmit) antenna spacing with respect to the carrier wavelength. The channel transfer function is therefore simplified to

$$H = \vec{a}_R(\theta_R) w \vec{a}_T^*(\theta_T) \tag{26}$$

where $w \sim \mathcal{N}_{\mathcal{C}}(0,1)$ is a complex Gaussian random variable. Without antenna selection, the FC-MRT/MRC is simply a beam

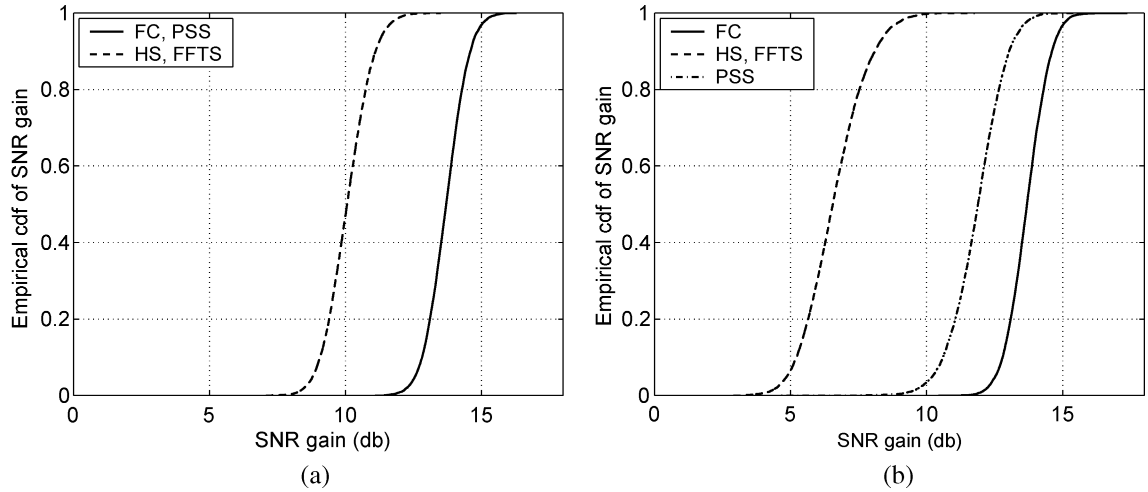


Fig. 4. CDF of the post SNR gain for the four schemes: FC-MRC/MRT, HS, FFTS and PSS under diversity transmission and i.i.d. Rayleigh faded MIMO channel with $t = r = 8$. (a) $L_r = L_R = 2$. (b) $L_T = L_R = 1$.

former in the direction of θ_T/θ_R. The distribution of the optimal SNR gain achieved is then

$$\gamma_{FC} = \frac{|\vec{a}_R^*(\theta_R)\vec{a}_R(\theta_R)w\vec{a}_T^*(\theta_T)\vec{a}_T(\theta_T)|^2}{rt} = rt|w|^2 \sim \Gamma(1, rt). \quad (27)$$

Here, $\Gamma(1, rt)$ denotes the Gamma distribution with parameters 1 and rt.[2]

It has been established in Section II-C that with $L_T, L_R \geq 2$, the PSS-MRT/MRC scheme can always deliver the same SNR gain as FC-MRT/MRC: $\gamma_{PSS}(L_T, L_R) = \gamma_{FC}$. It should be noted, however, that the optimum receiver weights $\vec{a}_R(\theta_R)$ in this case consisting of phase-shift-only entries, i.e., $\vec{a}_R(\theta_R) \in \mathcal{U}_{PSS}(1)$, are the transmit weights (modular a constant magnitude $1/\sqrt{t}$). Therefore, under such a channel condition the PSS can always achieve the full SNR gain

$$\gamma_{PSS}(L_T, L_R) = \gamma_{FC}, \quad 1 \leq L_T \leq t, \quad 1 \leq L_R \leq r.$$

To demonstrate the performance improvement of PSS over HS scheme, we note that

$$\begin{aligned}\gamma_{\mathrm{HS}} &= \max_{S_T \in \mathcal{S}_{L_T}} \max_{S_R \in \mathcal{S}_{L_R}} \lambda^2_{S_R \vec{a}_R(\theta_R) w \vec{a}_T(\theta_T) S_T, 1} \\ &= \max_{S_T \in \mathcal{S}_{L_T}} \max_{S_R \in \mathcal{S}_{L_R}} |w|^2 \, \|S_R \vec{a}_R(\theta_R)\|^2 \, \|\vec{a}_T^*(\theta_T) S_T^*\|^2 \\ &= L_T L_R |w|^2 \sim \Gamma(1, L_T L_R). \end{aligned} \quad (28)$$

HS-MRT/MRC can only obtain an average SNR gain of $L_T L_R$ (see Fig. 4).

As to the FFTS scheme, as an r-point FFT matrix is a beam former in the discrete directions, FFTS will deliver the same SNR gain as PSS and FC if the signal arrives in the directions on the FFT grid:

$$d_R \cos\theta_R = \frac{k}{r}, \quad k = 0, \ldots, r-1. \quad (29)$$

The performance at other directions will be degraded from γ_{FC}.

[2]The pdf of $X \sim \Gamma(\alpha, \beta)$ is $p_X(x) = (1/\beta^\alpha \Gamma(\alpha)) x^{\alpha-1} e^{-x/\beta}$ $(x > 0)$. The Γ function is $\Gamma(\alpha) = \int_0^\infty x^{\alpha-1} e^{-x} dx$. For a positive integer α, $\Gamma(\alpha) = (\alpha - 1)$.

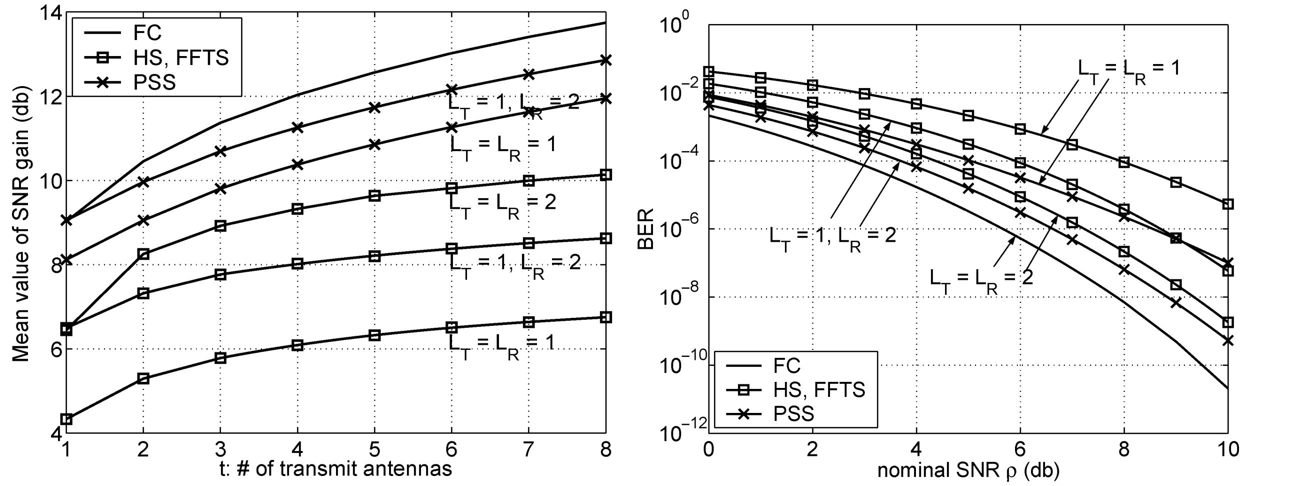
Simulation results are plotted and explained below with various channel conditions.

1) Independent Channel Fading

The Monte Carlo simulations are first conducted for uncorrelated MIMO channels with $C_T = I_t$ and $C_R = I_r$ (this corresponds to very large antenna spacings of d_T, d_R). The transfer function $H = W$ in this case becomes an i.i.d. Rayleigh fading channel. Fig. 5(a) demonstrates the average SNR gain versus the number of transmit antennas t with a fixed receiver setting $r = 8$. The PSS scheme outperforms HS and FFTS by far. The gap between PSS and HS/FFTS increases gradually with t: from 4 db at $t = 2$ to 5 db at $t = 8$ with $L_T = L_R = 1$. This is due to the fact that HS selection discards all the antennas that are not selected, therefore the power loss becomes more significant as more antennas are discarded with the increasing t; the PSS, on the other hand, still exploits the signals of all the t antennas through a simple RF chain reallocation (via the phase shifters). The average BER performance is plotted in Fig. 5(b) as a function of the SNR ρ at $t = r = 4$. The advantage of PSS over HS and FFTS is prominent: At $\rho = 10$ dB, the PSS has an average BER that is two orders of magnitude lower than for the other two.

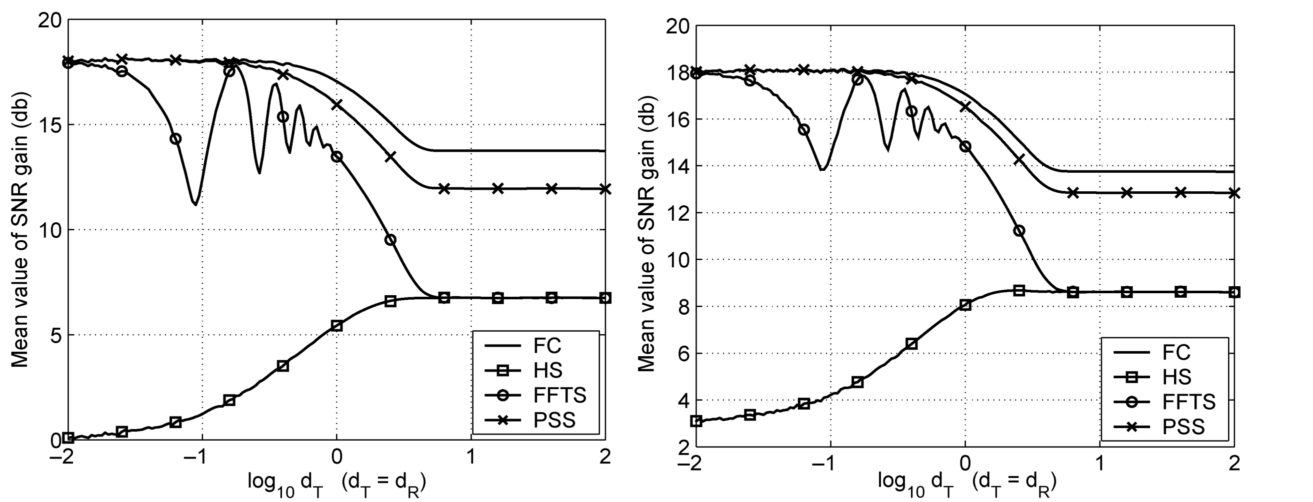
2) General Spatially Correlated Channels

For a spatially correlated MIMO channel of (25), the relative antenna spacings d_T and d_R reflect the degree of spatial correlation at the transmitter and receiver of the MIMO channel, respectively. Fig. 6 demonstrates the average SNR gain of the four schemes with respect to the relative antenna spacing. With small antenna spacings (e.g., $\log_{10} d_T = \log_{10} d_R < 0$), the channel coefficients present a strong correlation. In this case, HS performs considerably worse than the other schemes. The FFTS in general delivers an improvement, but the performance fluctuates in a cosine shape along the spacing axes. This phenomenon is due to the directional reception nature of FFTS, as illustrated in (29): as the beam forming direction is determined by $d_R \cos(\theta_R)$, the change in antenna



(a) (b)

Fig. 5. (a) SNR gain of the four schemes versus the number of transmit antennas t with various selection parameters L_T and L_R under diversity transmission. (b) BER performance of the four schemes versus the nominal SNR ρ with $t = r = 4$ under diversity transmission. In both figures, for $L_T = L_R = 2$, the PSS curve overlaps the FC curve.



(a) (b)

Fig. 6. Mean value of the post SNR gain for the four schemes versus the relative antenna spacing $d_T = d_R$ under diversity transmission. The channel parameters are $t = r = 8, \theta_T = \theta_R = \pi/4, \sigma_T = \sigma_R = 6°$. (a) $L_T = L_R = 2$. (b) $L_T = 1, L_R = 2$.

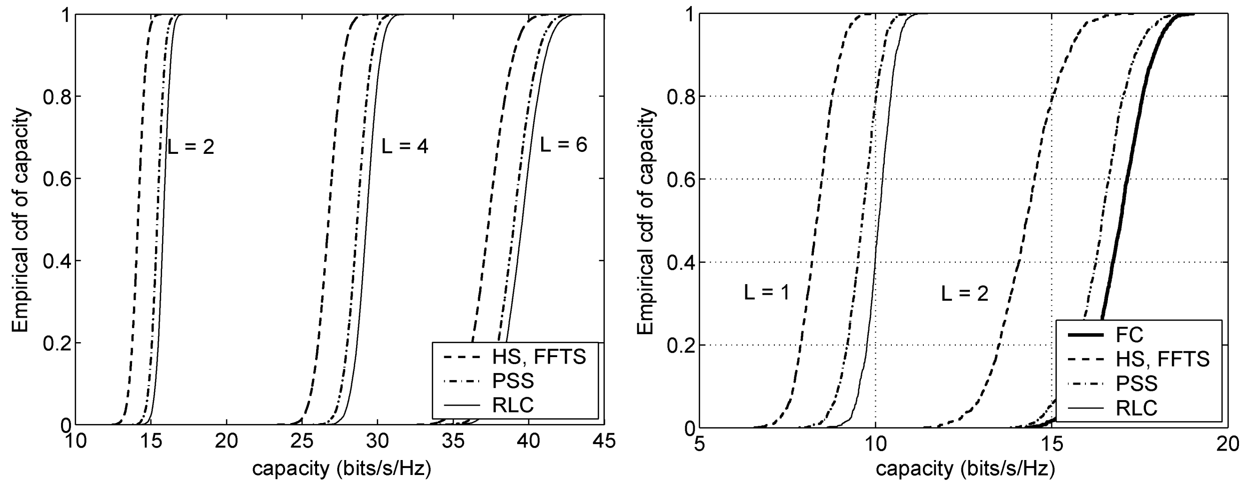
spacings leads to the variation in SNR gain. The fluctuation is reduced when allowing more selections, shown in the smoother plots of (b). Recall that in (27) and (28), for strongly correlated channels, the SNR gains for FC/PSS follow a distribution $\sim \Gamma(1, rt)$ with the mean value rt, and the HS has an average SNR gain of $L_t L_r$. This is further proved by the simulation results in the figure: when $\log_{10} d_T = \log_{10} d_R \leq -2$, for $L_T = L_R = 1$ in Fig. (a), we have the SNR gain of $10 \log_{10}(rt) \approx 18$ dB for FC/PSS, and $10 \log_{10}(L_t L_r) = 0$ dB for HS. The FFTS has the same gain with FC, as $d_R \cos \theta_R$ is near zero, which is on the FFT grid. With large antenna spacings (e.g., $\log_{10} d_T = \log_{10} d_R > 0$), the spatial correlation is weak and the fading is close to that of an independent Rayleigh channel. As a result, the FFTS gain drops dramatically, and approaches the HS values. In Fig. 6(b), the HS/FFTS (PSS) curves asymptotically deliver a 5 db (1 db) lower SNR gain than FC.

B. Spatial Multiplexing Transmission

We compared the capacity of PSS with HS and FFTS systems with spatial multiplexing. As indicated by Theorems 2 and 3, with multiplexing, there is an inevitable capacity loss for any linear selection schemes, and the loss is severe especially when $L \ll k_H$. For comparison we also provide the performance of the optimum RLC selection (or TRLC for the transceiver codesign; cf. Section III), as an upper bound of the linear selections.

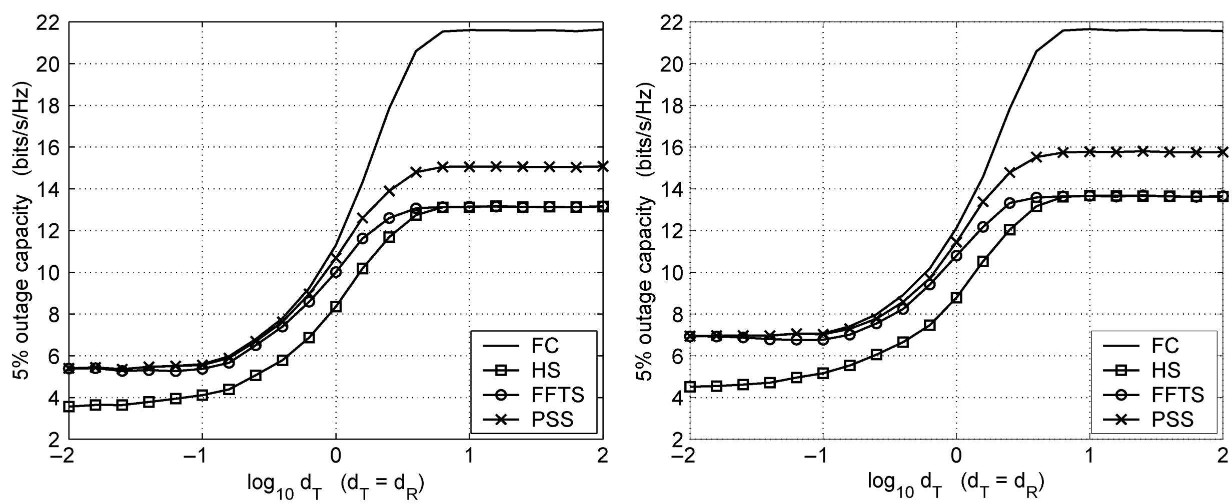
1) Independent Channel Fading

The capacity performances are plotted and compared in Fig. 7 for i.i.d. Rayleigh fading channels. With independent fading, the FFTS delivers the same performance as HS. With the antenna setting $t = r = 8$, all the selection schemes perform much worse than the full-complexity scheme when L is small, since the number of selected RF chains is much smaller than the rank of the channel. We see in Fig. 7(a) that the PSS is still above



(a) (b)

Fig. 7. Empirical cdf of the capacity for the antenna selection schemes in i.i.d. Rayleigh fading channels with spatial multiplexing. The SNR is $\rho = 20$ dB. Note that in (b), the RLC curve coincides with the FC curve when $L = 2$. (a) $t = r = 8$; Rx selection only. (b) $t = 2$, $r = 8$ joint Rx/Tx selection.



(a) (b)

Fig. 8. Outage capacity of 5% versus the relative antenna spacing $d_T = d_R$ under multiplexing transmission. The channel parameters are $t = 3$, $r = 8$, $\theta_T = \theta_R = \pi/6$, $\sigma_T = \sigma_R = 6°$, $\rho = 20$ dB. (a) $L = 2$; Rx selection only. (b) $L = 2$; joint Rx/Tx selection.

the other two and it is very close to the optimum performance of RLC, which performs the best linear selection. With the setting $t = 2$, $r = 8$, it is shown in Fig. 7(b) that HS/FFTS lose about 5 b/s/Hz in capacity while the PSS almost achieves the full channel capacity. When $L = 1$, the PSS has a 3 b/s/Hz advantage over the other two schemes. With CSI available at the transmitter, the joint selection always delivers a capacity improvement compared to the receiver-only design. The improvement is most significant when L is small compared with t and r; as in this case, it is critical to optimally allocate the transmitting power (rather than even distribution) to best utilize the system resource.

2) General Spatially Correlated Channels

Fig. 8 demonstrates the 5% outage capacity of the selection schemes with respect to the antenna spacings. In all the four figures, with small antenna spacings (strong correlation), the PSS and FFTS can approach the full channel capacity; when the spacings get large (weak correlation), FFTS delivers the same performance as HS, and PSS performs better than those. The PSS asymptotically improves the capacity by 3 and 2 b/s/Hz with receiver-only and joint transceiver selection, respectively. With any antenna spacings, the PSS is always close to the RLC (or TRLC) curve and is, therefore, near optimum among the linear selection schemes.

V. Conclusion

In this paper, we present a novel, soft antenna subset selection scheme for multiple antenna channels based on the joint design of RF chains and baseband processing. We address the antenna selection under two transmission strategies—diversity transmission and spatial multiplexing transmission, with focus on SNR gain and capacity, respectively. Transmitter and receiver selections are treated in duality. Standard antenna selection, as an L out of r switch at the front end of receiver, loses average signal power. This loss is most significant in highly correlated channels, but can also be significant with independent channel fading. Inserting an FFT operation before selection can help to improve the performance for strongly correlated channels with concentrated PAS but has no effect when the fading is independent. By using variable phase shifters adapting to the channel coefficients in the RF chains, the new scheme shows a prominent advantage in utilizing the multiple antenna diversity while incurring only a small hardware overhead. In either strongly correlated or independently fading MIMO channels, our approach always outperforms both the conventional pure antenna selection and FFT-based selection schemes. In particular, for the diversity transmission case with two branches selected, the new approach is able to achieve the same SNR gain as the full-complexity scheme involving all the antennas. Even when only one branch is selected, the performance is well above the conventional selection schemes. For spatial multiplexing, the capacity delivered by the PSS approach is very close to the optimum value within the linear framework. As the optimal (suboptimal) choices of the phase shifters are provided in closed form, the proposed antenna selection algorithm avoids the extensive computations of the conventional hard selection algorithms that require an exhaustive search for the best antenna subset. Computer experiments confirm that the proposed PSS scheme is an efficient way to exploit the multiple antenna diversity.

APPENDIX A
PROOF OF THEOREM 1

The necessary part is straightforward. If $L=1$, $\vec{u}$ reduces to a scalar. As Φ consists of only phase shifters, all the elements in the resulting vector $\vec{u}^*\Phi$ have the identical magnitude. It is then impossible to satisfy $\vec{u}^*\Phi=\vec{u}_{H,1}^*$, unless the channel has a special structure such that $\vec{u}_{H,1}$ has uniform amplitudes.

The sufficient part is verified constructively. When $L=2$, we denote the vector (matrix) elements in (6) as

$$\begin{aligned}\vec{u}^* &= [u_1 \quad u_2], \quad \Phi=\begin{bmatrix} e^{j\phi_{1,1}} & \dots & e^{j\phi_{1,r}} \\ e^{j\phi_{2,1}} & \dots & e^{j\phi_{2,r}}\end{bmatrix}\\ \vec{u}_{H,1}^* &= [\beta_{1,1}e^{-j\varphi_{1,1}} \quad \dots \quad \beta_{r,1}e^{-j\varphi_{r,1}}]\\ &\quad \beta_{i,1}\geq 0, \ \varphi_{i,1}\in\mathcal{R}. \end{aligned} \tag{30}$$

Given the optimal MRC vector $\vec{u}_{H,1}$ parameterized by ($\beta_{i,1}$, $\varphi_{i,1}$), the linear equation in (6) can be solved if and only if we can find $u_1, u_2 \in \mathcal{C}$ such that there exists r pairs of phases ($\phi_{1,i}$, $\phi_{2,i}$) satisfying

$$u_1 e^{j\phi_{1,i}} + u_2 e^{j\phi_{2,i}} = \beta_i e^{-j\varphi_{1,i}}, \quad \forall i=1,2,\dots,r. \tag{31}$$

From the triangle inequality, the norm of the composite complex number satisfies

$$||u_1|-|u_2|| \leq |u_1 e^{j\phi_{1,i}} + u_2 e^{j\phi_{2,i}}| \leq |u_1|+|u_2|. \tag{32}$$

Moreover, as $|u_1 e^{j\phi_{1,i}} + u_2 e^{j\phi_{2,i}}|$ is a continuous function of ($\phi_{1,i}$, $\phi_{2,i}$), the composite norm can achieve any number in the range of $[||u_1|-u_2||, |u_1|+|u_2|]$ with properly designed phases. Hence, the existence condition for solution to (31) is

$$||u_1|-|u_2|| \leq \beta_i \leq |u_1|+|u_2|, \quad \forall i=1,2,\dots,r. \tag{33}$$

Along this line, it is straightforward to show that the condition above is guaranteed by selecting

$$u_1=\frac{\beta_{\max}+\beta_{\min}}{2}, \quad u_2=\frac{\beta_{\max}-\beta_{\min}}{2} \tag{34}$$

where $\beta_{\max}=\max_{1\leq i\leq r}\beta_{i,1}$, $\beta_{\min}=\min_{1\leq i\leq r}\beta_{i,1}$ are the maximal and minimal norm among all the elements in the optimal weight vector $\vec{u}_{H,1}$. The solution to (31) with such a choice of $\vec{u}$ is obtained as

$$\begin{aligned}\phi_{1,i} &= -\varphi_{i,1}-\cos^{-1}\frac{\beta_{i,1}^2+\beta_{\max}\beta_{\min}}{\beta_{i,1}(\beta_{\max}+\beta_{\min})}\\ \phi_{2,i} &= -\varphi_{i,1}+\cos^{-1}\frac{\beta_{i,1}^2-\beta_{\max}\beta_{\min}}{\beta_{i,1}(\beta_{\max}-\beta_{\min})}.\end{aligned} \tag{35}$$

Thus, it is verified for $L=2$. Note that the solutions to (31) are not unique; any weights (u_1, u_2) satisfying (33) can lead to a valid set of phase design. The validity for $L>2$ is obvious as we can use the same parameter designs as for $L=2$ by setting the extra elements to be zero in vector $\vec{u}$.

APPENDIX B
PROOF OF THEOREM 2

Denote $R=BQ$ as a factorization of R, where B is an $L\times L$ matrix and Q $(L\times r)$ consists of orthonormal row vectors. The decomposition facilitates the simplification of (17):

$$C_{RLC}(L)=\max_{QQ^*=I_L}\log_2\det\left(I_t+\frac{\rho}{t}H^*Q^*QH\right). \tag{36}$$

Recall the singular value decomposition of H: $H=U_H\Sigma_H V_H^*$, which leads to $HH^*=U_H\Sigma_H\Sigma_H^*U_H^*$. Applying the determinant identity $\det(I+AB)=\det(I+BA)$, we obtain

$$\begin{aligned}C_{RLC}(L) &= \max_{QQ^*=I_L}\log_2\det\left(I_t+\frac{\rho}{t}HH^*Q^*Q\right)\\ &= \max_{QQ^*=I_L}\log_2\det\left(I_r+\frac{\rho}{t}U_H\Sigma_H\Sigma_H^*U_H^*Q^*Q\right)\\ &= \max_{\tilde{Q}\tilde{Q}^*=I_L}\log_2\det\left(I_L+\frac{\rho}{t}\tilde{Q}\Sigma_H\Sigma_H^*\tilde{Q}^*\right).\end{aligned} \tag{37}$$

The last equality is due to the fact that U_H is unitary and therefore $\tilde{Q}=QU_H$ shares the same orthonormal property of Q. As $\tilde{Q}$ is of size $L\times r$, the matrix $\tilde{Q}\Sigma_H\Sigma_H^*\tilde{Q}^*$ has at most L nonzero singular values. The capacity formula above is then equivalent to

$$C_{RLC}(L)=\max_{\tilde{Q}\tilde{Q}^*=I_L}\sum_{i=1}^{L}\log_2\left[1+\frac{\rho}{t}\lambda_{\tilde{Q}\Sigma_H\Sigma_H^*\tilde{Q}^*,i}\right]. \tag{38}$$

The matrix $\tilde{Q}\Sigma_H\Sigma_H^*\tilde{Q}^*$ is an $L\times L$ Hermitian matrix, and it is also the leading $L\times L$ principal submatrix of the $r\times r$ Hermitian matrix $\tilde{Q}'\Sigma_H\Sigma_H^*\tilde{Q}'^*$, where $\tilde{Q}'$ is $r\times r$ unitary and expanded from the rows vectors in $\tilde{Q}$. Then, from the interlacing property of the eigenvalues for Hermitian matrices [44, p. 411] [45, p. 103], it is straightforward to show that

$$\lambda_{\tilde{Q}\Sigma_H\Sigma_H^*\tilde{Q}^*,i}\leq\lambda_{\tilde{Q}'\Sigma_H\Sigma_H^*\tilde{Q}'^*,i}=\lambda_{H,i}^2, \quad 1\leq i\leq L. \tag{39}$$

From (38) and (39), we can prove that

$$C_{RLC}(L)\leq\sum_{i=1}^{L}\log_2\left(1+\frac{\rho}{t}\lambda_{H,i}^2\right). \tag{40}$$

For the achievability, it is obvious that the equality above can be obtained if, in (36), we let $Q = [\vec{u}_{H,1} \ \ldots \ \vec{u}_{H,L}]^*$. The optimal RLC matrix R could be such a Q multiplied by any rank-preserving matrix on the left. Therefore, we finally have the capacity result in the theorem.

Appendix C
Proof of Theorem 3

For simplicity, we make the assumption that $L \leq t$ in the following steps, although the same result remains valid when $L > t$. By directly applying the result in Theorem 2, when fixing T, the optimization over R is

$$C_{\mathrm{TRLC}}(L) = \max_{tr[TT^*]\leq t} \sum_{i=1}^{L} \log_2 \left(1 + \frac{\rho}{t}\lambda_{HT,i}^2\right) \quad (41)$$

$$= \max_{T\in\mathcal{C}^{t\times L}, tr[TT^*]\leq t} \log_2 \det\left(I_r + \frac{\rho}{t} HTT^*H^*\right) \quad (42)$$

and one of the optimal R is $R = [\vec{u}_{HT,1} \ \ldots \ \vec{u}_{HT,L}]^*$ [3]. From (42), we have

$$C_{\mathrm{TRLC}}(L) = \max_{T\in\mathcal{C}^{t\times L}, tr[TT^*]\leq t} \log_2 \det\left[I_L + \frac{\rho}{t} T^*\Sigma_H^*\Sigma_H T\right]. \quad (43)$$

Let $T = Q_T R_T$ be the standard QR factorization where Q_T is $t \times L$ unitary, and R_T is $L \times L$. The interlacing property of Hermitian matrix guarantees that $\lambda_{Q_T^*\Sigma_H^*\Sigma_H Q_{T},i} \leq \lambda_{\Sigma_H^*\Sigma_{H,I}} = \lambda_{H,i}^2$, $1 \leq i \leq L$. Therefore, we have

$$C_{\mathrm{TRLC}}(L) \leq \max_{R_T\in\mathcal{C}^{L\times L}, tr\left[R_T R_T^*\right]\leq t} \log_2 \det\left[I_L + \frac{\rho}{t} R_T^* \mathrm{diag}_{i=1}^L (\lambda_{H,i}^2) R_T\right]. \quad (44)$$

Formula (44) resembles that of the original channel capacity optimization problem in (21). Along the same line, (44) can be resolved via a similar water-filling strategy

$$R_T = \mathrm{diag}_{i=1}^L \left[\sqrt{\left(\mu - \frac{t}{\rho\lambda_{H,i}^2}\right)^+}\right]$$

where the constant μ satisfies $\sum_{i=1}^{L} (\mu - (t/\rho\lambda_{H,i}^2))^+ = t$. The achievability of (44) can be verified by directly substituting T in (42) with $T = [\vec{v}_{H,1} \ \ldots \ \vec{v}_{H,L}]R_T$.

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[3]The equivalence of (41) and (42) is verified by noting that for any matrix T satisfying $tr[TT^*] \leq t$, letting $HT = U_{HT}\Sigma_{HT}V_{HT}^*$ be the SVD of HT, we can construct a $t \times L$ matrix $\tilde{T} = T[\vec{v}_{HT,1} \ \ldots \ \vec{v}_{HT,L}]$. It is straightforward to show that the first L singular values of HT are the same as $H\tilde{T}$ with $\tilde{T}$ satisfying the constraint $tr[\tilde{T}\tilde{T}^*] \leq t$.

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