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感知無線網路接取技術及資源管理之研究–子計畫五：

協力式感知無線網路之頻譜偵測及接收機設計

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計畫主持人：吳卓諭 助理教授
共同主持人：王蒞君 教授
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摘要

為了實現高效能的感知無線網路，經由次要使用者之間的合作而獲得準確的頻譜偵測(亦稱為協力式頻譜偵測)被認為是不可缺少的架構。然而，這個技術不但要建立在高效率的實體層協力式通訊協定基礎上，也必須擁有完善的訊號收發機制。因此本研究計畫執行初期旨在開發一個高效能的協力式通訊架構以期有效提升連接品質。我們的研究方法是使用傳送端前置編碼(transmit-precoding)來探討此問題，明確來說我們特別提出一種結合來源端(source)與中繼端(relay)的前置編碼(joint source-relay precoding)的放大傳遞(amplify-and-forward)協力式通訊系統，最佳的前置編碼器是根據最小均方誤差準則(minimum mean square error criterion)所設計的。由於精確的最佳前置編碼器無法獲得，於是我們提出一種解析方法以獲得次佳的解析解。最後模擬的結果證實了我們所提出的方法比現有的中繼端前置編碼解法有更好的效能。

關鍵字：協力式頻譜偵測、協力式通訊、前置編碼、感知無線網路

Abstract

Accurate spectrum sensing via cooperation among secondary users (a.k.a. cooperative spectrum sensing) has been identified as an indispensable scheme to realize high-performance cognitve wireless networks. This technique, however, builds on efficient physical-layer cooperative communication protocols as well as powerful signal transmit-receive mechanisms. The initial-phase of this research project thus aims to develop a cooperative communication scheme for link quality enhancement. This problem is tackled via a transmit-precoding based approach. Specifically, we propose an amplify-and-forward cooperative system with joint source and relay precoding. The optimal source/relay precoders are designed based on the minimum mean square error criterion. While the exact joint optimal precoders are difficult to obtain, we propose an analytic approach to derive closed-form suboptimal solutions. Simulation study confirms the performance advantage of the proposed joint source-relay precoding scheme when compared with an existing relay-precoding solution.

Keywords: cooperative spectrum sensing; cooperative communication; precoding; cognitive radio network.

前言

為滿足高品質、高速率的多媒體通訊需求以及增加通訊使用者用戶，無線通訊實體層設計著重於如何增加通訊系統的頻譜使用率及如何提高通訊的鏈結品質與通訊容量。因此如何針對無線通訊的特性，設計合乎系統需求的通訊技術一直是新一代通訊標準所追求的目標。至目前為止，能達到高品質、高速率的關鍵無線傳輸技術即是在傳送與接收端配置多根天線，形成『多輸入多輸出系統』。然而，於實際通訊環境，因為使用者的手機體積限制，使的天線間的鏈結統計特性並未像學理所考慮的那樣完美，以致於多輸入多輸出系統效能表現出現瓶頸，因此，近年來協力式通訊(Cooperative communication)於感知無線網路系統中的應用已被廣泛地探討研究，其主要概念為利用中繼(Relay)節點幫忙傳送信號，使的系統能有較多且獨立鏈結路徑，藉以實現多輸入多輸出的系統概念，因此，這種系統也稱為虛擬多輸入多輸出系統。本計劃將針對協力式感知無線通訊系統做深入的分析研究，並提出新穎的協力式傳輸技術，以滿足未來無線通訊高品質、高速率的需求。

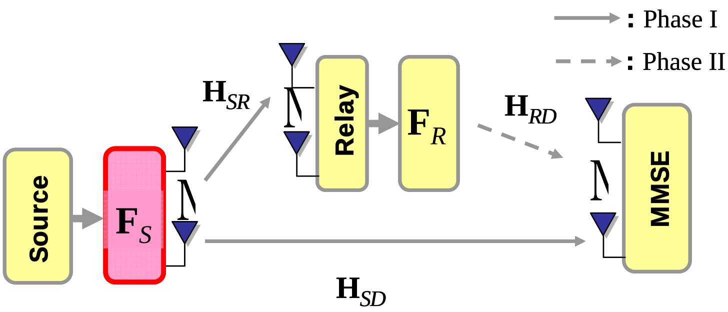
研究目的

通道衰減問題一直是無線通訊首當需要解決的問題。利用配置多根天線於多節點間，多輸入多輸出系統(Multiple-input multiple-output, MIMO)已被證明能提供空間多樣(Spatial diversity)來解決衰減所帶來的效能損失；此外，MIMO 亦能在不損失頻帶資源下，提供空間多工(Spatial multiplexing)來提升整體通道容量(Channel capacity)。然而，在點對點通訊的傳輸中，通道鏈結或稱為直接鏈結(Direct link)時常遭遇較差的大尺度衰減(Large scale fading)，這些時常會發生在如隧道內，或在大樓的通訊死角處。因此近年來協力式無線通訊系統被廣泛討論來解決這種直接鏈結不佳的情況，藉由配置中繼點(Relay)於強鏈結通道的地點，信號可以經由中繼點再傳至目的端，以減少因直接鏈結所產生的效能損失。而為了達到較高的傳輸率與通訊可靠度，近年來發展出一種在來源端(Source)、中繼端，與目的端(Destination)都配置多根天線的系統，稱為『多輸入多輸出中繼系統(MIMO Relay system)』，此系統不僅能效改善非協力式 MIMO 系統中的直接鏈結不佳的問題，並已被證明可以有效提升系統的多樣增益與多工增益。因此於本計劃中，我們將基於多輸入多輸出中繼系統中，利用通道鏈結的資訊，包含來源端-中繼端之通道、中繼端-目的端之通道及來源端-目的端之通道等訊息，提出新穎聯合式前置編碼(Joint precoders)設計，進一步改善此系統的通訊效能品質。不同於傳統 MIMO 系統中的前置編碼設計，在本計劃中我們將同時設計兩種前置編碼，分別稱為來源前置編碼(Source precoder)與中繼前置編碼(Relay precoder），合並設計以改良線性最小均方錯誤(Minimum mean-squared error, MMSE)接收機的效能表現。

文獻探討

對於多輸入多輸出中繼系統相關研究，目前著重於放大傳遞(Amplify-and-forward)的協力式傳輸協定架構的探討，這是因為相對於解碼傳遞(Decode-and-forward)，放大傳遞可以使中繼端有較低的操作複雜度與較低的延遲時間。對於此系統效能而言，目前已有相關論文分析多輸入多輸出中繼系統的容量上界(Capacity bound) [1]，其後，如同前置編碼相關技術發展於多輸入多輸出系統，相關前置編碼技術近年來也被發展於多輸入多輸出中繼系統[2]-[6]。其中，中繼端的前置編碼首先被設計來增加接收端的傳輸率[2], [3]或增加鏈結品質[4], [5], [6]。在這些文獻中，都只考量中繼點的前置編碼來設計並改善多輸入多輸出中繼系統的效能，所以來源端的自由度尚未利用。此外，現有的這類系統的中繼前置編碼設計，為了簡化最佳化過程的困難度，大部分[2], [4]-[5]都只考慮中繼連結(Relay link），所謂的中

繼鏈結就是來源端至中繼端與中繼端至目的端的通道，所以直接鏈結的通訊資源並未考慮。因此，有別於現有文獻的前置編碼設計，本計劃將補強現有文獻所尚未考慮的系統資源，我們將考量來源端與中繼端的傳輸自由度，並根據此系統所提供的鏈結資源，包含直接鏈結與中繼鏈結，合併設計來源端與中繼端的前置編碼，使的接收機的最小均方錯誤接收機能有顯著的效能改善(如圖一)。



圖一: 多輸入多輸出中繼系統之聯合試前置編碼系統。

研究方法

在這研究子題中，我們將引入來源端的前置編碼於多輸入多輸出中繼系統中，並考量直接鏈結的信號，重新推導線性解碼矩陣(Decoding matrix)和接收端的最小均方(MSE)效能。必然的，這個最小均方的值一定為當時的通道鏈結以及兩個前置編碼器的函數，因此，我們將設計這兩個前置編碼器使的接收端的最小均方效能最好，並同時滿足來源端與中繼端的最大功率限制，如下所示:

$$\begin{aligned}
&\min_{\mathbf{F}_S,\mathbf{F}_R} MSE\left(\mathbf{F}_S,\mathbf{F}_R\right)\\
&s.t.\\
&tr\left(\mathbf{F}_S E\left[\mathbf{s}\mathbf{s}^H\right]\mathbf{F}_S^H\right)\le P_{S,T}\\
&tr\left(\mathbf{F}_R E\left[\mathbf{y}_R\mathbf{y}_R^H\right]\mathbf{F}_R^H\right)\le P_{R,T}
\end{aligned},$$

其中 $\mathbf{F}_S$ 與 $\mathbf{F}_R$ 分別代表來源端與中繼端的前置編碼；$\mathbf{y}_R$ 代表中繼端所收到的信號，$\mathbf{s}$ 代表傳送信號；$P_{S,T}$ 與 $P_{R,T}$ 分別代表來源端與中繼端的最大功率限制。進一步化簡，我們可得以下最佳化問題。

$$\begin{aligned}
&\min_{\mathbf{F}_S,\mathbf{F}_R} tr\Big\{\Big(\sigma_s^{-2}\mathbf{I}_L+\sigma_{n,d}^{-2}\mathbf{F}_S^H\mathbf{H}_{SD}^H\mathbf{H}_{SD}\mathbf{F}_S+\\
&+\mathbf{F}_S^H\mathbf{H}_{SR}^H\mathbf{F}_R^H\mathbf{H}_{RD}^H\left(\sigma_{n,r}^2\mathbf{H}_{RD}\mathbf{F}_R\mathbf{F}_R^H\mathbf{H}_{RD}^H+\sigma_{n,d}^2\mathbf{I}_M\right)^{-1}\mathbf{H}_{RD}\mathbf{F}_R\mathbf{H}_{SR}\mathbf{F}_S\Big)^{-1}\Big\}\\
&s.t.\\
&\sigma_s^2 tr\left\{\mathbf{F}_S^H\mathbf{F}_S\right\}\le P_{S,T}\\
&tr\left\{\mathbf{F}_R\left(\sigma_{n,r}^2\mathbf{I}_R+\sigma_s^2\mathbf{H}_{SR}\mathbf{F}_S\mathbf{F}_S^H\mathbf{H}_{SR}^H\right)\mathbf{F}_R^H\right\}\le P_{R,T},
\end{aligned}$$

其中，$\mathbf{H}_{SD}$ 、$\mathbf{H}_{SR}$ 與 $\mathbf{H}_{RD}$ 分別為來源端至目的端、來源端至中繼端與中繼端至目的端之通道矩陣； $\sigma_{n,r}^2$ 與 $\sigma_{n,d}^2$ 分別代表中繼端與目的端的雜訊功率。

因為最小均方接收器引入了反矩陣的計算操做，這會使的成本含數(Cost function)表示

成前置編碼的高度非線性函數，因此將提升最佳化過程的困難度。所以，於本計劃中，我們將提出新穎的信號處裡方法，主要概念為提出受限制的前置編碼架構，如下所示:

$$\mathbf{F}_S = \mathbf{V}_{sr} \sum\nolimits_s ; \ \mathbf{F}_R = \mathbf{V}_{rd} \sum\nolimits_r \mathbf{U}_{sr}^H .$$

其中只有對角矩陣 $\sum_s = diag\left(\sigma_{s,1},\cdots,\sigma_{s,L}\right)$ 與 $\sum_r = diag\left(\sigma_{r,1},\cdots,\sigma_{r,L}\right)$ 需要設計；$\mathbf{H}_{SD} = \mathbf{U}_{sd}\Sigma_{sd}\mathbf{V}_{sd}^H$；$\mathbf{H}_{SR} = \mathbf{U}_{sr}\Sigma_{sr}\mathbf{V}_{sr}^H$；$\mathbf{H}_{RD} = \mathbf{U}_{rd}\Sigma_{rd}\mathbf{V}_{rd}^H$。利用此前置編碼架構與對角化概念，我們可以推導出最小均方上限

$$tr\left(\mathbf{E}\right) \le \sum_{i=1}^{L} \frac{1}{\sigma_s^{-2} + \dfrac{\sigma_{s,i}^2\sigma_{r,i}^2\sigma_{sr,i}^2\sigma_{rd,i}^2}{\sigma_{n,r}^2\sigma_{r,i}^2\sigma_{rd,i}^2 + \sigma_{n,d}^2} + \sigma_{n,d}^{-2}\dfrac{\sigma_{s,i}^2}{\mathbf{B}^{-1}(i,i)}},$$

其中，$\mathbf{B} = \mathbf{V}_{sr}^H\mathbf{V}_{sd}\Sigma_{sd}^H\Sigma_{sd}\mathbf{V}_{sd}^H\mathbf{V}_{sr}$。利用所推導的最小均方上限，將使的原本需用矩陣的最佳化，轉換成存數的最佳化，如下所示:

$$\begin{aligned}
&\min_{\sigma_{s,i},\sigma_{r,i}} \sum_{i=1}^{L} \frac{1}{\sigma_s^{-2} + \dfrac{\sigma_{s,i}^2\sigma_{r,i}^2\sigma_{sr,i}^2\sigma_{rd,i}^2}{\sigma_{n,r}^2\sigma_{r,i}^2\sigma_{rd,i}^2 + \sigma_{n,d}^2} + \sigma_{n,d}^{-2}\dfrac{\sigma_{s,i}^2}{\mathbf{B}^{-1}(i,i)}} \\
&s.t. \\
&\sum_{i=1}^{L} p_{r,i}\left(\sigma_{n,r}^2 + \sigma_s^2 p_{s,i}\sigma_{sr,i}^2\right) \le P_{R,T} \\
&\sigma_s^2\sum_{i=1}^{L} p_{s,i} \le P_{S,T}, p_{s,i} = \sigma_{s,i}^2 \ge 0, p_{r,i} = \sigma_{r,i}^2 \ge 0 \quad \forall i
\end{aligned}$$

相較於原來矩陣佳化問題，這裡所提出的最佳化將降低運算複雜度，進而計算出聯合式前置編碼的解析解，如下所示。

$$p_{r,i} = \frac{1}{\left(\sigma_s^2 p_{s,i}\sigma_{sr,i}^2 + \sigma_{n,r}^2\right)} \times \left[\frac{\mu_r\sqrt{p_{s,i}}\sigma_{n,d}\sigma_{sr,i}\sigma_{rd,i}\left(\sigma_s^2 p_{s,i}\sigma_{sr,i}^2 + \sigma_{n,r}^2\right)^{1/2}}{\sigma_{rd,i}^2\left(\sigma_{n,r}^2\left(\sigma_s^{-2} + \sigma_{n,d}^{-2}p_{s,i}\left(\mathbf{B}^{-1}(i,i)\right)^{-1}\right) + p_{s,i}\sigma_{sr,i}^2\right)} - \frac{\left(\sigma_s^2 p_{s,i}\sigma_{sr,i}^2 + \sigma_n^2\right)\sigma_{n,r}^2\left(\sigma_s^{-2} + \sigma_{n,d}^{-2}p_{s,i}\left(\mathbf{B}^{-1}(i,i)\right)^{-1}\right)}{\sigma_{rd,i}^2\left(\sigma_{n,r}^2\left(\sigma_s^{-2} + \sigma_{n,d}^{-2}p_{s,i}\left(\mathbf{B}^{-1}(i,i)\right)^{-1}\right) + \sigma_n^{-2}p_{s,i}\sigma_{sr,i}^2\right)}\right]^+ ;$$

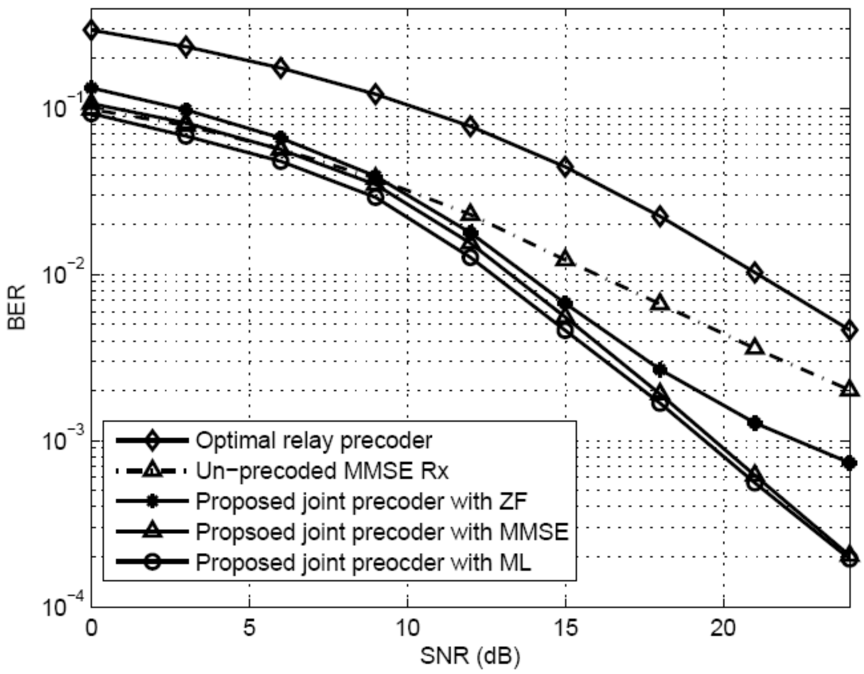
$$p_{s,i} = \left[\frac{\mu_s\sqrt{\beta_i} - \sigma_s^2\left(\sigma_{n,d}^2 + p_{r,i}\sigma_{n,r}^2\sigma_{rd,i}^2\right)}{\sigma_{n,d}^{-2}\left(\left(\mathbf{B}^{-1}(i,i)\right)^{-1}\left(\sigma_{n,d}^2 + p_{r,i}\sigma_{n,r}^2\sigma_{rd,i}^2\right) + p_{r,i}\sigma_{sr,i}^2\sigma_{rd,i}^2\right)}\right]^+ ,$$

其中 $\beta_i = \left(\sigma_{n,d}^2 + p_{r,i}\sigma_{n,r}^2\sigma_{rd,i}^2\right)\left(\sigma_{n,d}^{-2}\left(\mathbf{B}^{-1}(i,i)\right)^{-1}\left(\sigma_{n,d}^2 + p_{r,i}\sigma_{n,r}^2\sigma_{rd,i}^2\right) + \sigma_n^{-2}p_{r,i}\sigma_{sr,i}^2\sigma_{rd,i}^2\right)$； μ_r 與 μ_s 是用來設計滿足中繼端與來源端的功率，詳細推導請參考[7]。

結果與討論

圖二主要說明了所提出的方法與既有文獻[4],[5]的效能比較。於圖中，我們固定直接

鏈結的(Signal-to-noise ratio, SNR)訊雜比，並改變中繼鏈結的訊雜比。最上面的那條線代表只有用中繼端的前置編碼與只考慮直接鏈結的效能表現[4],[5]；最下面那三條分別表示本計劃所提出的前置編碼器應用於不同的接收機的效能；虛線代表沒有任何前置編碼器的效能表現。如圖所示，有考量聯合式前置編碼的效能優於只用中繼前置編碼，這是因為我們在此計畫中不但用了中繼端的前置編碼器，而且還考量直接鏈結資源並加上了來源端的前置編碼器聯合處裡，但是因為在此所設計的前置編碼只針對最小均方錯誤接收機做設計，所以所提出的聯合式前置編碼應用於最大相似接收機的效能表現有限。



圖二：本計劃所提出的聯合式前置編碼與其他方法的位元錯誤率效能比較。

成果自評

雖然計劃執行未滿一年，本研究目前以有相當具體的成果，相關研究結果已投稿到 IEEE Transactions on Wireless Communications 期刊接受審查中，同時也已經先藉由會議論文的方式發表於 IEEE Wireless Communications and Networking conference 2009 [7] (全文詳見附檔)，可見本研究今年度的成果已具有相當優異且具體的學術價值。

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Joint Source/Relay Precoder Design in Amplify-and-Forward Relay Systems Using an MMSE Criterion

Fan-Shuo Tseng, Wen-Rong Wu, Jwo-Yuh Wu
Department of Communication Engineering
National Chiao-Tung University
Hsinchu Taiwan 300
stron.cm93g@nctu.edu.tw; wrwu@faculty.nctu.edu.tw; jywu@cc.nctu.edu.tw

***Abstract*—This paper addresses the joint source/relay precoder design problem in amplify-and-forward (AF) cooperative communication systems where multiple antennas are equipped at the source, the relay, and the destination. Existing solutions to the problem only consider the relay link and, thus, do not fully exploit all the available link resource. Using a minimum-mean-squared-error (MMSE) criterion, we propose a joint precoder design method, taking both the direct and relay links into account. It is shown that the MMSE is a highly nonlinear function of the precoder matrices, and a direct minimization is not feasible. To facilitate analysis, we propose to design the precoders toward first diagonalizing the MSE matrix of the relay link. This imposes certain structural constraints on both precoders that allow us to derive an analytically tractable MSE upper bound. By conducting minimization with respect to this upper bound, the solution can be obtained by an iterative water-filling technique. Simulations show that the proposed design can significantly enhance the performance of MIMO AF cooperative systems.**

I. INTRODUCTION

Multi-input-multi-output (MIMO) techniques are known to be capable of providing an extra dimension for capacity/ diversity enhancement for wireless communications [1]-[5]. However, in practical applications, it may be difficult to place multiple antennas (with sufficient spacing) in a wireless unit due to the size or other constraints. Recently, cooperative communication (CC) is developed to overcome this problem [6]-[12]. In the physical layer two well-known cooperative protocols are amplify-and-forward (AF) and decode-and-forward (DF). In the AF protocol, the relays retransmit the received signals with only signal amplification. In contrast, the DF scheme decodes the received signal at the relay, re-encodes the information bits, and retransmits the resultant signal to the destination. In general, the computational complexity the DF protocol is larger than that of AF.

Precoder designs in AF-based MIMO-CC systems have been recently addressed in the literature, either for capacity enhancement [9], [10], or for signal decoding performance improvement via minimization of the symbol mean square errors [11], [12]. A common feature of these works is that the system scenario focuses on signal precoding only at the relay node. The works [9] and [11] even neglect the effect of the direct link in order to ease analysis. To the best of our knowledge, the general precoder designs for AF-based MIMO-CC systems that consider joint source-and-relay precoding and take into account all the available link resource have not yet been addressed in the literature.

This paper aims to study the joint source/relay precoding problem for AF-based MIMO-CC systems from an MMSE perspective, taking both the direct and relay links into consideration. It is shown that the MMSE function in this general system scenario is a highly nonlinear function of the source and relay precoders, and direct minimization with the cost function is not feasible. To overcome the problem, we propose to design the precoders toward first diagonalizing the MSE matrix of the relay link. This imposes certain structural constraints on both precoders that allow us to derive a tractable MSE upper bound. Minimization with this upper bound, instead of the original MMSE, then becomes feasible. The resultant solutions can be obtained via an iterative water-filling technique. Simulations show that the proposed method can significantly improve the BER performance as compared to the non-precoded system.

II. SYSTEM MODEL

A. Precoders for AF Systems

We consider a typical three-node cooperative AF system in which multiple antennas are placed at the source, relay and destination. Under this system configuration, signals can be transmitted from the source to the destination (direct link), and from the source to the relay, and then to the destination (relay link). To avoid interference, orthogonal channels can be used for both links, as in the time-division-duplexing scheme [9], [10]. Let N, R, and M denote the number of antennas at the source, the relay, and the destination, respectively, and assume that all channels are flat-fading. In the first phase, the received signals at the destination and the relay from the source can be expressed as

$$\mathbf{y}_{D,1} = \mathbf{H}_{SD}\mathbf{F}_S\mathbf{s} + \mathbf{n}_{D,1} \tag{1}$$

and

$$\mathbf{y}_R = \mathbf{H}_{SR}\mathbf{F}_S\mathbf{s} + \mathbf{n}_R, \tag{2}$$

respectively, where $\mathbf{s} \in \mathbb{C}^{L\times 1}$ is the transmitted signal vector, $\mathbf{F}_S \in \mathbb{C}^{N\times L}$ is the precoding matrix at source, $\mathbf{H}_{SR} \in \mathbb{C}^{R\times N}$

is the channel matrix between source and relay, $\mathbf{H}_{SD} \in \mathbb{C}^{M\times N}$ is the channel matrix between source to destination, $\mathbf{n}_{D,1} \in \mathbb{C}^{M\times 1}$ is the first-phase received noise vector at destination, and $\mathbf{n}_R \in \mathbb{C}^{R\times 1}$ is the received noise vector at relay. Here, we assume $L \le N, R, M$ to provide sufficient degrees of freedom for transmission.

In the second phase, the relay transmits the received signal with another precoding matrix. Thus, the received signal at the destination can be expressed as

$$\mathbf{y}_{D,2} = \mathbf{H}_{RD}\mathbf{F}_R\mathbf{y}_R + \mathbf{n}_{D,2}, \tag{3}$$

where $\mathbf{F}_R \in \mathbb{C}^{R\times R}$ is the precoding matrix at relay, $\mathbf{H}_{RD} \in \mathbb{C}^{M\times R}$ is the channel matrix between the relay and destination, and $\mathbf{n}_{D,2} \in \mathbb{C}^{M\times 1}$ is the second-phase received noise vector at destination.

B. Linear MMSE Receiver at Destination

The received signal vectors $\mathbf{y}_{D,1}$ and $\mathbf{y}_{D,2}$ can be combined into a single vector as

$$\mathbf{y}_D := \begin{bmatrix}\mathbf{y}_{D,1}\\ \mathbf{y}_{D,2}\end{bmatrix} = \mathbf{Hs} + \mathbf{w}, \tag{4}$$

where

$$\mathbf{H} = \begin{bmatrix}\mathbf{H}_{SD}\mathbf{F}_S\\ \mathbf{H}_{RD}\mathbf{F}_R\mathbf{H}_{SR}\mathbf{F}_S\end{bmatrix},\ \mathbf{w} = \begin{bmatrix}\mathbf{n}_{D,1}\\ \mathbf{H}_{RD}\mathbf{F}_R\mathbf{n}_R + \mathbf{n}_{D,2}\end{bmatrix}. \tag{5}$$

Here, $\mathbf{H}$ is the combined channel matrix, and $\mathbf{w}$ is the combined noise vector at destination. Note that noise received at the relay is amplified by the relay-to-destination link. Also note that the precoder design problem is a joint transceiver problem, that is, a different receiver will yield a different precoder. Here, we propose to use the linear minimum mean-squared-error (MMSE) receiver at the destination. Then, the mean-squared-error (MSE), denoted as J, is given by

$$J = E\left\{\left\|\mathbf{G}\mathbf{y}_D - \mathbf{s}\right\|^2\right\}. \tag{6}$$

Minimizing (6), we can obtain [2]

$$\mathbf{G}_{opt} = \mathbf{R}_s\mathbf{H}^H\left(\mathbf{H}\mathbf{R}_s\mathbf{H}^H + \mathbf{R}_w\right)^{-1}, \tag{7}$$

where $\mathbf{G}_{opt}$ is the optimum solution, $\mathbf{R}_w = E\left[\mathbf{w}\mathbf{w}^H\right]$ is the covariance matrix of the combined noise vector $\mathbf{w}$, and $\mathbf{R}_s = E\left[\mathbf{s}\mathbf{s}^H\right]$ is that of the signal vector. Substituting (7) into (6), we can have the MMSE, denoted as J_{min}, as:

$$J_{\min} = tr\left\{\left(\mathbf{R}_s^{-1} + \mathbf{H}^H\mathbf{R}_w^{-1}\mathbf{H}\right)^{-1}\right\}. \tag{8}$$

Assume that

1) $\mathbf{R}_{\mathbf{n}_{D,1}} = E\left[\mathbf{n}_{D,1}\mathbf{n}_{D,1}^H\right] = \sigma_n^2\mathbf{I}_M$, $\mathbf{R}_{\mathbf{n}_{D,2}} = E\left[\mathbf{n}_{D,2}\mathbf{n}_{D,2}^H\right] = \sigma_n^2\mathbf{I}_M$, *and* $\mathbf{R}_R = E\left[\mathbf{n}_R\mathbf{n}_R^H\right] = \sigma_n^2\mathbf{I}_R$, where σ_n^2 is the noise variance of each vector element.

2) The elements of the signal vectors are i.i.d. with zero-mean and covariance $\mathbf{R}_s = \sigma_s^2\mathbf{I}_L$, where σ_s^2 is the transmitted symbol power of each element. Direct manipulations show

$$J_{\min} = tr\{\mathbf{E}\} = tr\left\{\underbrace{\left(\sigma_s^{-2}\mathbf{I}_L + \mathbf{E}_S + \mathbf{E}_R\right)^{-1}}_{:=\mathbf{E}}\right\}, \tag{9}$$

where

$$\mathbf{E}_S = \sigma_n^{-2}\mathbf{F}_S^H\mathbf{H}_{SD}^H\mathbf{H}_{SD}\mathbf{F}_S, \tag{10}$$

and

$$\mathbf{E}_R = \sigma_n^{-2}\mathbf{F}_S^H\mathbf{H}_{SR}^H\mathbf{F}_R^H\mathbf{H}_{RD}^H\left(\mathbf{H}_{RD}\mathbf{F}_R\mathbf{F}_R^H\mathbf{H}_{RD}^H + \mathbf{I}_M\right)^{-1}\times \mathbf{H}_{RD}\mathbf{F}_R\mathbf{H}_{SR}\mathbf{F}_S. \tag{11}$$

We note that $\mathbf{E}_S$ and $\mathbf{E}_R$ can be regarded as the MSE components due to the direct and relay links, respectively. It is also noteworthy that, by ignoring the direct link and adopting a precoder only at relay, the MMSE designing criterion was also considered in [11], [12] in the AF MIMO system. Here, we further incorporate the precoder at source and consider both the direct and relay link signals to enhance the performance of the considered system.

C. Problem Formulation

Based on (9)-(11) the MMSE-based joint source/relay precoder design problem can be formulated as

$$\begin{aligned}&\min_{\mathbf{F}_S,\mathbf{F}_R} tr\left\{\underbrace{\left(\sigma_s^2\mathbf{I}_L + \mathbf{E}_S + \mathbf{E}_R\right)^{-1}}_{:=\mathbf{E}}\right\} \quad s.t\\ &tr\left\{\mathbf{F}_R\left(\sigma_n^2\mathbf{I}_R + \sigma_s^2\mathbf{H}_{SR}\mathbf{F}_S\mathbf{F}_S^H\mathbf{H}_{SR}^H\right)\mathbf{F}_R^H\right\} \le P_{R,T}\\ &\sigma_s^2 tr\left\{\mathbf{F}_S\mathbf{F}_S^H\right\} \le P_{S,T},\end{aligned} \tag{12}$$

where $P_{S,T}$ and $P_{R,T}$ are the maximal available power at source and relay, respectively. The inequalities in (12) indicates that the designed precoders have to satisfy the transmit power constraints. Since the cost function in (12) is highly nonlinear in $\mathbf{F}_S$ and $\mathbf{F}_R$, it appears quite difficult to directly seek for the optimal precoding matrices. In the next section, we propose an analytical approach to find suboptimal solutions.

III. Joint Source/Relay Precoder Design

A. Proposed Approach

It is observed that, if the error matrix $\mathbf{E}$ in (9) can be diagonalized, the trace operation can be easily conducted, and whole problem can be greatly simplified. Motivated by these findings, we propose to seek for certain $\mathbf{F}_S$ and $\mathbf{F}_R$ with which $\mathbf{E}$ is as close to being diagonal as possible. To do that, let us first perform the singular value decomposition (SVD) for the link channel matrices as:

$$\mathbf{H}_{SD} = \mathbf{U}_{sd} \Sigma_{sd} \mathbf{V}_{sd}^H ; \tag{13}$$

$$\mathbf{H}_{SR} = \mathbf{U}_{sr} \Sigma_{sr} \mathbf{V}_{sr}^H ; \tag{14}$$

$$\mathbf{H}_{RD} = \mathbf{U}_{rd} \Sigma_{rd} \mathbf{V}_{rd}^H , \tag{15}$$

where $\mathbf{U}_{sd} \in \mathbb{C}^{M\times M}$, $\mathbf{U}_{sr} \in \mathbb{C}^{R\times R}$, and $\mathbf{U}_{rd} \in \mathbb{C}^{M\times M}$ are left singular vector matrices of $\mathbf{H}_{SD}$, $\mathbf{H}_{SR}$, and $\mathbf{H}_{RD}$, respectively; $\mathbf{V}_{sd}^H \in \mathbb{C}^{N\times N}$, $\mathbf{V}_{sr}^H \in \mathbb{C}^{N\times N}$, and $\mathbf{V}_{rd}^H \in \mathbb{C}^{R\times R}$ are the right singular vector matrices of $\mathbf{H}_{SD}$, $\mathbf{H}_{SR}$, and $\mathbf{H}_{RD}$, respectively. In addition, we define $\sigma_{sd,i}$, $\sigma_{sr,i}$, and $\sigma_{rd,i}$ as the ith diagonal element of $\Sigma_{sd} \in \mathbb{R}^{M\times N}$, $\Sigma_{sr} \in \mathbb{R}^{R\times N}$, and $\Sigma_{rd} \in \mathbb{R}^{M\times R}$, respectively.

From (9), it is seen that the error matrix $\mathbf{E}$ can be diagonal if we can choose $\mathbf{F}_S$ and $\mathbf{F}_R$ to simultaneously diagonalize $\mathbf{E}_S$ and $\mathbf{E}_R$. Such a task, however, appears quite difficult to achieve mainly because $\mathbf{E}_R$ depends on the relay precoder $\mathbf{F}_R$ through matrix inversion $\left(\mathbf{H}_{RD}\mathbf{F}_R\mathbf{F}_R^H\mathbf{H}_{RD}^H + \mathbf{I}_M\right)^{-1}$. This thus motivates us to first choose $\mathbf{F}_R$ to diagonalize $\mathbf{H}_{RD}\mathbf{F}_R\mathbf{F}_R^H\mathbf{H}_{RD}^H + \mathbf{I}_M$ so that the inverse can be easily tracked. Such an approach, though suboptimal, will considerably simplify the analysis; more importantly, it allows us to derive an tractable MSE upper bound which will lead to a water-filling based solution. Based on (13)-(15), such $\mathbf{F}_S$ and $\mathbf{F}_R$ can be shown to be

$$\mathbf{F}_R = \mathbf{V}_{rd} \Sigma_r \mathbf{U}_{sr}^H , \tag{16}$$

and

$$\mathbf{F}_S = \mathbf{V}_{sr} \Sigma_s . \tag{17}$$

Based on (16) and (17), the cost function (9) can be further rearranged as

$$tr\{\mathbf{E}\} = tr\left\{\left(\sigma_s^{-2}\mathbf{I}_L + \underbrace{\sigma_n^{-2}\Sigma_s^H \mathbf{V}^H \Sigma_{sd}^H \Sigma_{sd} \mathbf{V} \Sigma_s}_{=\mathbf{E}_S} + \underbrace{\sigma_n^{-2}\Sigma_s^H \Sigma_{sr}^H \Sigma_r^H \Sigma_{rd}^H \left(\Sigma_{rd}\Sigma_r^2\Sigma_{rd}^H + \mathbf{I}_M\right)^{-1}\Sigma_{rd}\Sigma_r\Sigma_{sr}\Sigma_s}_{=\mathbf{E}_R}\right)^{-1}\right\}, \tag{18}$$

where $\mathbf{V} = \mathbf{V}_{sd}^H\mathbf{V}_{sr}$ is a constant matrix depending on the channels. We make the following key observations regarding the alternative MSE expression (18):

- Since (18) is obtained by the particular precoding matrices $\mathbf{F}_R$ in (16) and $\mathbf{F}_S$ in (17), it serves as an upper bound of true minimal MSE.
- Compared with the original MSE formula (9), the expression (18) is more appealing because the unknowns involved are Σ_r and Σ_s, which are diagonal matrices and are more amenable to handle.
- The matrix $\mathbf{E}_S$ cannot be diagonalized. However, starting from (18) and exploiting the diagonal nature of $\mathbf{E}_R$, we can derive a more tractable MSE upper bound that will be used as the design cost function, as shown next.

To proceed, let us use the matrix inversion lemma [13] to rewrite (18) as

$$\begin{aligned} tr(\mathbf{E}) &= tr\left(\left[\underbrace{\left(\sigma_s^{-2}\mathbf{I}_L + \mathbf{E}_R\right)}_{:=\mathbf{A}} + \Sigma_s^H \underbrace{\left(\sigma_n^{-2}\mathbf{V}^H\Sigma_{sd}^H\Sigma_{sd}\mathbf{V}\right)}_{:=\mathbf{B}}\Sigma_s\right]^{-1}\right) \\ &= tr\left(\mathbf{A}^{-1}\right) - tr\left(\mathbf{A}^{-1}\Sigma_s^H\left(\mathbf{B}^{-1} + \Sigma_s\mathbf{A}^{-1}\Sigma_s^H\right)^{-1}\Sigma_s\mathbf{A}^{-1}\right). \end{aligned} \tag{19}$$

Based on (19), the desired MSE upper bound can be obtained with the aid of the next lemma (We have proved the lemma [16], and omitted the details here).

Lemma: Let $\mathbf{D}_1$ and $\mathbf{D}_2$ be diagonal matrices, with the diagonal entries of $\mathbf{D}_2$ being positive. Then for any positive definite matrix $\mathbf{X}$, we have

$$tr\left(\mathbf{D}_1^H\left(\mathbf{X}+\mathbf{D}_2\right)^{-1}\mathbf{D}_1\right) \geq tr\left(\mathbf{D}_1^H\left(diag\{\mathbf{X}\}+\mathbf{D}_2\right)^{-1}\mathbf{D}_1\right), \tag{20}$$

where $diag\{\mathbf{X}\}$ is obtained from $\mathbf{X}$ by setting its off-diagonal entries to be zero. □

By the lemma, it follows

$$\begin{aligned} &tr\left(\mathbf{A}^{-1}\Sigma_s^H\left(\mathbf{B}^{-1}+\Sigma_s\mathbf{A}^{-1}\Sigma_s^H\right)^{-1}\Sigma_s\mathbf{A}^{-1}\right) \geq \\ &tr\left(\mathbf{A}^{-1}\Sigma_s^H\left(diag\left(\mathbf{B}^{-1}\right)+\Sigma_s\mathbf{A}^{-1}\Sigma_s^H\right)^{-1}\Sigma_s\mathbf{A}^{-1}\right) \end{aligned}. \tag{21}$$

Based on (20) and (21), we reach the following key result

$$\begin{aligned} &tr(\mathbf{E}) \leq \\ &tr\left(\mathbf{A}^{-1}\right) - tr\left(\mathbf{A}^{-1}\Sigma_s^H\left(diag\left(\mathbf{B}^{-1}\right)+\Sigma_s\mathbf{A}^{-1}\Sigma_s^H\right)^{-1}\Sigma_s\mathbf{A}^{-1}\right) \\ &= \sum_{i=1}^{L}\frac{1}{\sigma_s^{-2}+\sigma_n^{-2}\dfrac{\sigma_{s,i}^2\sigma_{r,i}^2\sigma_{sr,i}^2\sigma_{rd,i}^2}{\sigma_{r,i}^2\sigma_{rd,i}^2+1}+\sigma_n^{-2}\dfrac{\sigma_{s,i}^2}{\mathbf{B}^{-1}(i,i)}}, \end{aligned} \tag{22}$$

where the last equality follows from some straightforward manipulations. Compared with the origin MSE in (9), the upper bound (22) admits a simple rational form and is analytically tractable. Hence, we propose to design the precoders by minimizing the MSE upper bound (22). By setting $p_{s,i} = \sigma_{s,i}^2$, $p_{r,i} = \sigma_{r,i}^2$ in (22) and with (16)-(17), the optimization problem is reformulated as

$$\min_{p_i, p_{r,i}, i=1,\cdots,L} \sum_{i=1}^{L} \frac{1}{\sigma_s^{-2} + \sigma_n^{-2} \frac{p_{s,i} p_{r,i} \sigma_{sr,i}^2 \sigma_{rd,i}^2}{p_{r,i}\sigma_{rd,i}^2 + 1} + \sigma_n^{-2} \frac{p_{s,i}}{\mathbf{B}^{-1}(i,i)}}$$

s.t.

$$tr\left\{\Sigma_r \left(\sigma_n^2 \mathbf{I}_R + \sigma_s^2 \Sigma_{sr} \Sigma_s \Sigma_s^H \Sigma_{sr}^H\right) \Sigma_r^H\right\}$$

$$= \sum_{i=1}^{L} p_{r,i}\left(\sigma_n^2 + \sigma_s^2 p_{s,i} \sigma_{sr,i}^2\right) \le P_{R,T}, \quad p_{r,i} \ge 0.$$

$$\sigma_s^2 tr\left\{\Sigma_s \Sigma_s^H\right\} = \sum_{i=1}^{L} p_{s,i} \le P_{S,T}, \quad p_{s,i} \ge 0. \tag{23}$$

B. Iterative Waterfilling

Since the optimization in (23) is not convex, there is no global optimum. We thus propose an iterative waterfilling approach to find a suboptimal solution. For this we can resort to the standard Lagrange technique. By solving the resultant set of KKT conditions we have (details omitted due to space limitation)

$$p_{r,i} = \frac{1}{\left(\sigma_s^2 p_{s,i} \sigma_{sr,i}^2 + \sigma_n^2\right)} \times \left[\frac{\mu_r \sqrt{p_{s,i}} \sigma_{sr,i} \sigma_{rd,i} \left(\sigma_s^2 p_{s,i} \sigma_{sr,i}^2 + \sigma_n^2\right)^{1/2}}{\sigma_{rd,i}^2 \left(\sigma_s^{-2} + \sigma_n^{-2} p_{s,i} \left(\mathbf{B}^{-1}(i,i)\right)^{-1} + \sigma_n^{-2} p_{s,i} \sigma_{sr,i}^2\right)} - \frac{\left(\sigma_s^2 p_{s,i} \sigma_{sr,i}^2 + \sigma_n^2\right)\left(\sigma_s^{-2} + \sigma_n^{-2} p_{s,i} \left(\mathbf{B}^{-1}(i,i)\right)^{-1}\right)}{\sigma_{rd,i}^2 \left(\sigma_s^{-2} + \sigma_n^{-2} p_{s,i} \left(\mathbf{B}^{-1}(i,i)\right)^{-1} + \sigma_n^{-2} p_{s,i} \sigma_{sr,i}^2\right)}\right]^+, \tag{24}$$

where, $[y]^+ = \max[0, y]$ and μ_r is the water level which should be chosen to satisfied the power constraint at relay. Also, solution for $p_{s,i}$ can be expressed as

$$p_{s,i} = \left[\frac{\mu_s \sqrt{\beta_i} - \sigma_s^2\left(1 + p_{r,i}\sigma_{rd,i}^2\right)}{\sigma_n^{-2}\left(\left(\mathbf{B}^{-1}(i,i)\right)^{-1}\left(1 + p_{r,i}\sigma_{rd,i}^2\right) + p_{r,i}\sigma_{sr,i}^2 \sigma_{rd,i}^2\right)}\right]^+, \tag{25}$$

where μ_s is the water level which is chosen to meet the total power constraint $\sum_{i=1}^{L} p_{s,i} = P_{S,T}$ at source node, and

$$\beta_i = \left(1 + p_{r,i}\sigma_{rd,i}^2\right) \times \left(\sigma_n^{-2}\left(\mathbf{B}^{-1}(i,i)\right)^{-1}\left(1 + p_{r,i}\sigma_{rd,i}^2\right) + \sigma_n^{-2} p_{r,i} \sigma_{sr,i}^2 \sigma_{rd,i}^2\right). \tag{26}$$

Equation (24) and (25) show that there is an interdependence between $p_{s,i}$ and $p_{r,i}$. Thus, we can use an iterative approach similar to [17] to find a local optimum, and substitute the results into (16) and (17) to obtain the final solution.

IV. APPLICATIONS

In this section, we conduct simulations to evaluate the performance of the proposed precoding scheme in a one-hop MIMO relay system (relay only system), and a general MIMO relay system. We assume that perfect synchronization can be obtained, and exact channel state information is available at all nodes. Also, the modulation scheme is QPSK. Let the elements of each channel matrix be i.i.d. complex Gaussian random variable with zero mean and unit variance. Also let $P_{S,T} = P_{R,T} = P/2$ and the SNR be defined as P/σ_n^2.

A. Two-Hop MIMO Relay Systems

In this scenario, the channel condition in the direct link is very poor so that the destination only receives the signal from the relay link. We consider the case that $N = R = M = L = 4$. Fig. 1 shows the performance comparison of the proposed precoding scheme with the MMSE receiver and the un-precoded systems with the ZF and MMSE receivers. From the figure, we can see that the proposed system outperforms the un-precoded system by 6 dB at $BER = 10^{-2}$ (for the MMSE receiver).

B. General MIMO Relay Systems

In this scenario, the symmetric cooperative case with $N = R = M = L = 2$ and the case with $N = R = M = 4$, $L = 2$, are compared. Fig. 2 shows the performance comparison for the proposed and un-precoded schemes. The result is similar to what we have observed in the previous case; the proposed precoding scheme yields improved performance. Also, as we can see, the performance of the case with $N = R = M = L = 2$ is worst than that when $N = R = M = 4$, $L = 2$. This is because the later case provides more degrees of freedom for effective signal precoding.

V. CONCLUSIONS

In this paper, we propose a joint source/relay precoder design method for AF MIMO cooperative systems. At the destination the MMSE receiver is implemented, and precoders are designed to minimize the MMSE. Since the MMSE is a complicated function of precoder matrices, a direct minimization is not feasible. To solve the problem we propose to design the precoders by first diagonalizing the MSE matrix of the relay link. This imposes certain structural constraints on both precoders which allow us to derive a tractable MSE upper bound. Since the upper bound has a simple expression, minimization with the upper bound becomes feasible. Resorting to the KKT optimality conditions, we finally derive an iterative waterfilling algorithm to obtain a suboptimum solution. We then consider the performance of the proposed scheme in two-hop MIMO and general MIMO systems. Simulations show that, in all the considered system scenarios, the proposed scheme can have significant performance improvement compared to the non-precoded systems.

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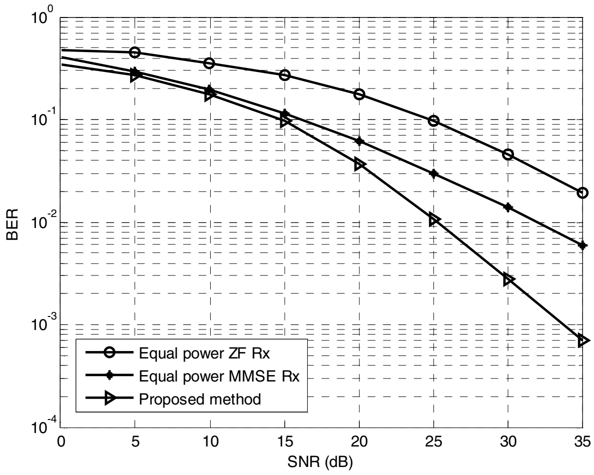

Figure 1. BER perfroamce comparison for precoded and non-precoded schemes in the two-hop system.

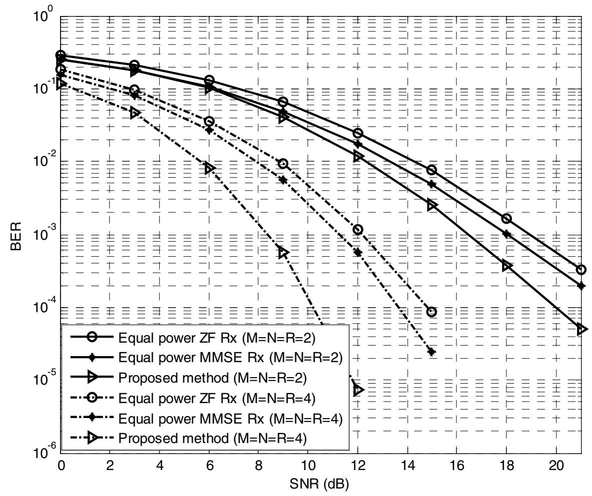

Figure 2. BER perfroamce comparison for precoded and non-precoded schemes in MIMO AF CC system.

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