An Efficient Detector for Combined Space–Time Coding and Layered Processing

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***Abstract*—Group layered space–time architecture (GLST) combines space–time block coding and layered space–time processing, where the transmit stream is partitioned into different groups, and in each group, space–time block coding is applied. In the traditional receiver of GLST, group detection is applied first to suppress the interference from other groups, and then decoding is performed for the desired group. In this letter, a novel detector is proposed in which the entire groups are decoded first, and then group detection is performed next. Theoretical analysis will demonstrate that the new detector can achieve a significant capacity gain compared with the traditional one. Simulation results will further show that the proposed detector can obtain at least 4 dB gain at a frame-error rate of 10^{-2}, for instance.**

***Index Terms*—Group detection, layered space–time (LST), multiple-input/multiple-output (MIMO) systems, outage capacity, space–time block coding.**

I. Introduction

MULTIPLE-INPUT/multiple-output (MIMO) systems can provide both diversity gain and multiplexing gain [1]. Most existing MIMO techniques aim at achieving either maximum diversity gain or maximum multiplexing gain. For example, space–time codes (STC) (including space–time block codes (STBC) [2], [3] and space–time trellis codes (STTC) [4]) are carefully designed to achieve the full diversity order, but no multiplexing gain can be obtained. Layered space–time (LST), such as V-BLAST [5], can achieve maximum multiplexing gain, but with a very low diversity gain. Group layered space–time architecture (GLST) has been proposed to achieve a better tradeoff between multiplexing gain and diversity gain, in which the transmit stream is partitioned into different groups, and in each group, STC is applied [6]. It can be regarded as a combination of STC and LST. This combination of coding and array processing at the receiver provides much better multiplexing gain than STC with lower decoding complexity, while at the same time achieving a much higher diversity gain than LST.

One possible approach in the detection of GLST, for a given group, is to suppress signals transmitted from other groups of antennas by virtue of a group detector first, and then perform space–time decoding for the desired group. This detector, which we refer to as the Type I detector, was proposed in [6], and adopted widely later [7], [8]. However, it is not necessarily the most effective one. We found that when an STBC is applied inside groups at the transmitter of GLST, an alternative detector is to perform decoding first and then do group detection. Specifically, for a GLST system with m transmit and n receive antennas during T time slots, the linear nature of STBC can be exploited to obtain an equivalent $Tn \times m$ channel. Group detection can then be applied to this equivalent channel, instead of the original $n \times m$ one. It can be seen that after decoding, the receive dimensions increase from n to Tn, and thus, better performance can be achieved by group detection. Considering that the entire performance is limited by the step of group detection, it can be expected that the latter detector, which is referred to as the Type II detector, should have superior performance over the Type I detector.

In this letter, a novel detector for GLST, i.e., a Type II detector, is proposed. We analyze the mutual information achieved by the Type II detector and compare it with that of the Type I detector. Performance is evaluated based upon the outage capacity and frame-error rate (FER). It will be shown that the Type II detector can achieve at least 4 b/s/Hz capacity gain over the Type I detector. Besides, with the Type II detector, much higher diversity gain can be obtained. At a FER of 10^{-2}, at least 4 dB gain can be achieved, for instance. This substantial gain is, however, achieved at a cost of a complexity increase, as we will show in Section IV.

This letter is organized as follows. The system model is provided in Section II. In Section III, we review the Type I detector first, and then present the details of the Type II detector. The mutual information of both detectors is also derived. Simulation results are given in Section IV, where performance is evaluated based upon the outage capacity and FER. A complexity comparison of both detectors will be also provided. Finally, Section V summarizes and concludes this letter.

II. System Model

We consider a wireless link with m transmit and n receive antennas. Assume that the channel remains constant within a block of L symbols. Let h_{ij} denote the complex path gain from transmit antenna j to receive antenna i, which is modeled as samples of independent complex Gaussian random variables with mean zero and variance 0.5 per dimension. We also assume perfect channel knowledge at the receiver side only, through the use of training sequences.

As shown in Fig. 1, all the m transmit antennas are partitioned into G groups, respectively, comprising $m_1, m_2, \ldots, m_G$ antennas with $\sum_{i=1}^{G} m_i = m$. A block of input symbols $\{b_i\}_{i=1,\ldots,K}$ with length K is divided into G groups, $\mathcal{G}_1, \mathcal{G}_2, \ldots, \mathcal{G}_G$, and in each group, $\mathbf{b}_i = \left[b_{i,1}, b_{i,2}, \ldots, b_{i,|\mathcal{G}_i|}\right]'$, $i = 1, \ldots, G$ is then encoded by a component space–time block code STBC_i, associated with m_i transmit antennas. Assume

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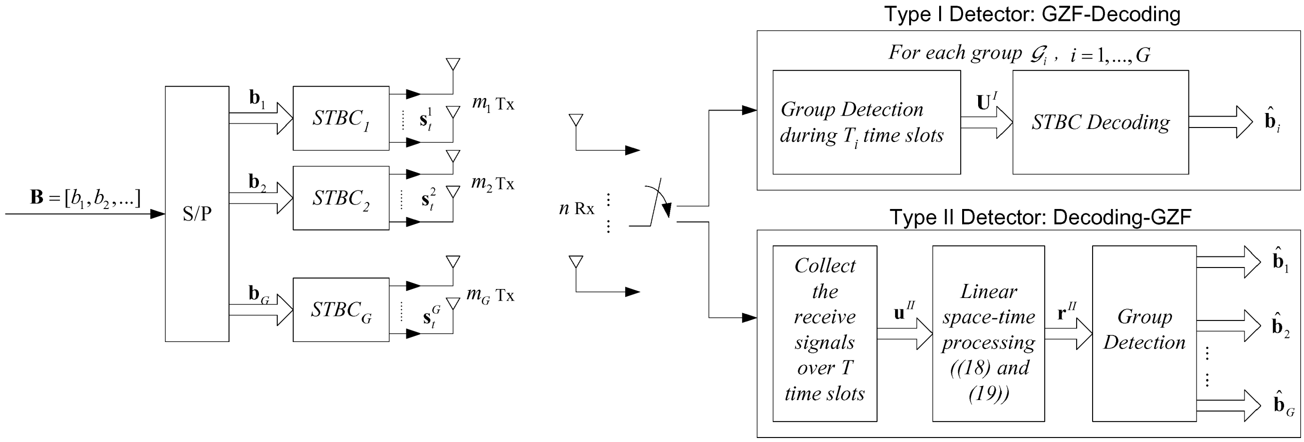

Fig. 1. Block diagram of GLST.

that the component codes STBC_i, $i=1,\ldots,G$ are with a code length T_i, respectively, and that the block length T is selected to be the minimum common multiple of $\{T_1,\ldots,T_G\}$.[1] Then, the output $m\times T$ codeword matrix $\mathbf{X}$ over a block of T symbol intervals can be written as

$$\mathbf{X}=\begin{bmatrix}\mathbf{s}_1^1 & \cdots & \mathbf{s}_T^1\\ \vdots & \ddots & \vdots\\ \mathbf{s}_1^G & \cdots & \mathbf{s}_T^G\end{bmatrix}=\begin{bmatrix}\mathbf{S}_1\\ \vdots\\ \mathbf{S}_G\end{bmatrix} \tag{1}$$

where $\mathbf{S}_i=[\mathbf{s}_1^i,\ldots,\mathbf{s}_T^i]=[\mathbf{A}_{i1}\mathbf{b}_i,\ldots,\mathbf{A}_{iT}\mathbf{b}_i]+[\mathbf{B}_{i1}\mathbf{b}_i^*,\ldots,\mathbf{B}_{iT}\mathbf{b}_i^*]$ is the $m_i\times T$ codeword matrix of group $\mathcal{G}_i$, $i=1,\ldots,G$, with $(\cdot)^*$ denoting the conjugate operator, and $\mathbf{A}_{ij}$, $\mathbf{B}_{ij}$ are constant coefficient matrices in $\mathcal{R}^{m_i\times|\mathcal{G}_i|}$, $j=1,\ldots,T_i$.

The $n\times T$ discrete received complex signal $\mathbf{Y}$ over T time slots can now be written as

$$\begin{aligned}\mathbf{Y}&=\sqrt{\frac{\text{SNR}}{m}}\mathbf{H}\mathbf{X}+\mathbf{Z}\\ &=\sqrt{\frac{\text{SNR}}{m}}\left\{[\mathbf{H}_1,\ldots,\mathbf{H}_G]\cdot\begin{bmatrix}\mathbf{A}_{11}\mathbf{b}_1 & \cdots & \mathbf{A}_{1T}\mathbf{b}_1\\ \vdots & \ddots & \vdots\\ \mathbf{A}_{G1}\mathbf{b}_G & \cdots & \mathbf{A}_{GT}\mathbf{b}_G\end{bmatrix}\right.\\ &\left.\quad+[\mathbf{H}_1,\ldots,\mathbf{H}_G]\cdot\begin{bmatrix}\mathbf{B}_{11}\mathbf{b}_1^* & \cdots & \mathbf{B}_{1T}\mathbf{b}_1^*\\ \vdots & \ddots & \vdots\\ \mathbf{B}_{G1}\mathbf{b}_G^* & \cdots & \mathbf{B}_{GT}\mathbf{b}_G^*\end{bmatrix}\right\}+\mathbf{Z}\end{aligned} \tag{2}$$

where the additive white noise $\mathbf{Z}$ has independent and identically distributed (i.i.d.) entries z_{it}, $i=1,\ldots,n$, $t=1,\ldots,T$, which are all Gaussian random variables with mean zero and unit variance. $\mathbf{H}_i$ is the $n\times m_i$ channel matrix of group $\mathcal{G}_i$, $i=1,\ldots,G$. Likewise, SNR is the average signal-to-noise ratio at each receive antenna.

III. GLST Detectors

An intuitive approach for the detection of GLST is to extract the signals of the desired group via group detection and then perform space–time decoding. Another alternative is to do space–time decoding first, and then perform group detection. The above detectors are referred to as the Type I detector and Type II detector, respectively. In this section, we will present the details of the Type II detector and analyze the mutual information of both detectors. We begin by reviewing the Type I detector.

A. Type I Detector

When the Type I detector is adopted, group detection is performed first. In this letter, group zero-forcing (GZF) is assumed to be deployed. Assume that group $\mathcal{G}_i$ is to be detected. Then, the interference from the other groups $\mathcal{G}_1,\ldots,\mathcal{G}_{i-1},\mathcal{G}_{i+1},\ldots,\mathcal{G}_G$ should be nulled out using an orthogonal projection. To obtain the projection matrix, we partition $\mathbf{H}$ into $\mathbf{H}=[\mathbf{H}_i,\mathbf{H}_{\bar{i}}]$, where $\mathbf{H}_{\bar{i}}$ includes the columns of $\mathbf{H}$ corresponding to all the groups except $\mathcal{G}_i$. The projection matrix $\mathbf{P}_i$ is then defined as [9], [11]

$$\mathbf{P}_i=\mathbf{I}_n-\mathbf{H}_{\bar{i}}(\mathbf{H}_{\bar{i}}^+\mathbf{H}_{\bar{i}})^{-1}\mathbf{H}_{\bar{i}}^+ \tag{3}$$

where $(\cdot)^+$ denotes the conjugate and transpose operator. Obviously, it is required that $n\geq m-m_i+1$ to guarantee a valid $\mathbf{P}_i$.

Using the linear transformation, $\mathbf{W}_i=\mathbf{H}_i^+\mathbf{P}_i$, on the received signal $\mathbf{Y}$, we get

$$\mathbf{U}^{\text{I}}=\mathbf{W}_i\cdot\mathbf{Y}=\sqrt{\frac{\text{SNR}}{m}}\mathbf{Q}_i^{-1}\mathbf{S}_i+\mathbf{V}^{\text{I}} \tag{4}$$

where

$$\mathbf{Q}_i^{-1}=\mathbf{H}_i^+\mathbf{P}_i\mathbf{H}_i. \tag{5}$$

It turns out that $\mathbf{Q}_i=\left[(\mathbf{H}^+\mathbf{H})^{-1}\right]_{m_i\times m_i}$ is the $m_i\times m_i$ diagonal submatrix of $(\mathbf{H}^+\mathbf{H})^{-1}$. That is, it is composed of the intersection elements of rows from $\sum_{j=1}^{i-1}m_j+1$ to $\sum_{j=1}^{i}m_j$ and columns from $\sum_{j=1}^{i-1}m_j+1$ to $\sum_{j=1}^{i}m_j$ in $(\mathbf{H}^+\mathbf{H})^{-1}$. The covariance of the noise $\mathbf{V}^{\text{I}}$ is then given by $\text{Cov}\left(\mathbf{v}_t^I\right)=\mathbf{Q}_i^{-1}$, where $\mathbf{v}_t^I$ is the tth column vector of $\mathbf{V}^I$, $t=1,\ldots,T_i$.

[10] has proposed an algorithm for constructing any orthogonal STBC that guarantees that $\mathbf{b}$ and $\mathbf{b}^*$ will not appear in the same time slot t. Under this assumption, (4) is equivalent to

$$\mathbf{r}^{\text{I}}=\sqrt{\frac{\text{SNR}}{m}}\boldsymbol{\mathcal{H}}_i\cdot\mathbf{b}_i+\mathbf{z}^{\text{I}} \tag{6}$$

[1]Assume that T is less than L. That is, the channel remains constant within T symbols.

where $\mathbf{r}^{\mathrm{I}} = \left[\left(\tilde{\mathbf{y}}_1^{\mathrm{I}}\right)', \ldots, \left(\tilde{\mathbf{y}}_{T_i}^{\mathrm{I}}\right)'\right]'$, $(\cdot)'$ represents the transpose operator, and

$$\tilde{\mathbf{y}}_j^I = \begin{cases} \mathbf{u}_j^I, & \text{if } \mathbf{B}_{ij} = \mathbf{0}_{m_i \times |\mathcal{G}_i|} \\ \left(\mathbf{u}_j^I\right)^*, & \text{if } \mathbf{A}_{ij} = \mathbf{0}_{m_i \times |\mathcal{G}_i|} \end{cases} \quad (7)$$

with $\mathbf{u}_j^I$ denoting the jth column vector of $\mathbf{U}^I$, $j = 1, \ldots, T_i$. For any group $\mathcal{G}_i$, $i = 1, \ldots, G$, its corresponding subchannel matrix $\mathcal{H}_i$ is given by

$$\mathcal{H}_i = \begin{bmatrix} \mathbf{Q}_i^{-1}\mathbf{A}_{i1} + \left(\mathbf{Q}_i^{-1}\right)^* \mathbf{B}_{i1}^* \\ \vdots \\ \mathbf{Q}_i^{-1}\mathbf{A}_{iT_i} + \left(\mathbf{Q}_i^{-1}\right)^* \mathbf{B}_{iT_i}^* \end{bmatrix}. \quad (8)$$

Likewise, $\mathrm{Cov}(\mathbf{z}^{\mathrm{I}}) = \mathrm{diag}(\tilde{\mathbf{Q}}_j)$, i.e., $\mathrm{Cov}(\mathbf{z}^{\mathrm{I}})$ is made up of T_i diagonal submatrix $\tilde{\mathbf{Q}}_j$, where

$$\tilde{\mathbf{Q}}_j = \begin{cases} \mathbf{Q}_i^{-1}, & \text{if } \mathbf{B}_{ij} = \mathbf{0}_{m_i \times |\mathcal{G}_i|} \\ \left(\mathbf{Q}_i^{-1}\right)^*, & \text{if } \mathbf{A}_{ij} = \mathbf{0}_{m_i \times |\mathcal{G}_i|} \end{cases}, \quad j = 1, \ldots, T_i. \quad (9)$$

From the property of $\{\mathbf{A}_{ij}\}_{j=1,\ldots,T_i}$ and $\{\mathbf{B}_{ij}\}_{j=1,\ldots,T_i}$, we have

$$\mathcal{H}_i^+ \cdot \mathcal{H}_i = \mathrm{trace}\left(\mathbf{Q}_i^{-1} \cdot \left(\mathbf{Q}_i^{-1}\right)^+\right) \cdot \mathbf{I}_{|\mathcal{G}_i|} \quad (10)$$

where $\mathbf{I}_{|\mathcal{G}_i|}$ denotes a $|\mathcal{G}_i| \times |\mathcal{G}_i|$ identity matrix. Let $\lambda = \mathrm{trace}(\mathbf{Q}_i^{-1} \cdot (\mathbf{Q}_i^{-1})^+)$, then

$$\hat{\mathbf{r}}^{\mathrm{I}} = \mathcal{H}_i^+ \mathbf{r}^{\mathrm{I}} = \sqrt{\frac{\mathrm{SNR}}{m}} \lambda \mathbf{b}_i + \hat{\mathbf{z}}^{\mathrm{I}}, \quad (11)$$

where

$$\mathrm{Cov}\left(\hat{\mathbf{z}}^{\mathrm{I}}\right) = \mathcal{H}_i^+ \mathrm{Cov}\left(\mathbf{z}^{\mathrm{I}}\right) \mathcal{H}_i = \mathcal{H}_i^+ \mathrm{diag}\left(\tilde{\mathbf{Q}}_j\right) \mathcal{H}_i. \quad (12)$$

By combining (11) and (12), the mutual information between the input $\mathbf{b}_i$ and output $\hat{\mathbf{r}}^{\mathrm{I}}$ over a channel given by $\sqrt{\mathrm{SNR}/m}\lambda \mathbf{I}_{|\mathcal{G}_i|}$ can be obtained as

$$\begin{aligned} &I\left(\mathbf{b}_i; \hat{\mathbf{r}}^{\mathrm{I}} | \mathbf{H} = \lambda \mathbf{I}_{|\mathcal{G}_i|}\right) \\ &\quad = \log\det\left(\frac{\mathcal{H}_i^+ \mathrm{diag}\left(\tilde{\mathbf{Q}}_j\right)\mathcal{H}_i + \frac{\mathrm{SNR}}{m} \cdot \lambda^2 \mathbf{I}_{|\mathcal{G}_i|}}{\mathcal{H}_i^+ \mathrm{diag}\left(\tilde{\mathbf{Q}}_j\right)\mathcal{H}_i}\right). \quad (13) \end{aligned}$$

In [7], another linear transformation $\mathbf{W}_i = \mathbf{D}_i^+ \mathbf{P}_i$, instead of $\mathbf{H}_i^+ \mathbf{P}_i$, is applied on $\mathbf{Y}$, where $\mathbf{D}_i$ is an $n \times (n - m + m_i)$ matrix which satisfies $\mathbf{D}_i \mathbf{D}_i^+ = \mathbf{P}_i$ and $\mathbf{D}_i^+ \mathbf{D}_i = \mathbf{I}_{n-m+m_i}$. In this case, we have

$$\mathbf{U}^{Ii} = \sqrt{\frac{\mathrm{SNR}}{m}} \mathbf{D}_i^+ \mathbf{H}_i \mathbf{S}_i + \mathbf{V}^{Ii} \quad (14)$$

where $\mathrm{Cov}(\mathbf{v}_t^{Ii}) = \mathbf{I}_{n-m+m_i}$, for $t = 1, \ldots, T_i$. Then, after decoding

$$\hat{\mathbf{r}}^{Ii} = \sqrt{\frac{\mathrm{SNR}}{m}} \mathrm{trace}\left(\mathbf{Q}_i^{-1}\right) \cdot \mathbf{b}_i + \hat{\mathbf{z}}^{Ii} \quad (15)$$

where $\mathrm{Cov}(\hat{\mathbf{z}}^{Ii}) = \mathrm{trace}(\mathbf{Q}_i^{-1}) \cdot \mathbf{I}_{|\mathcal{G}_i|}$. Therefore, the mutual information can be given as follows:

$$\begin{aligned} &I\left(\mathbf{b}_i; \hat{\mathbf{r}}^{Ii} | \mathbf{H} = \mathrm{trace}\left(\mathbf{Q}_i^{-1}\right) \cdot \mathbf{I}_{|\mathcal{G}_i|}\right) \\ &\qquad = |\mathcal{G}_i| \log\left(1 + \frac{\mathrm{SNR}}{m} \cdot \mathrm{trace}\left(\mathbf{Q}_i^{-1}\right)\right). \quad (16) \end{aligned}$$

It can be seen that this detector has a similar structure to that of the Type I detector. That is, group detection is performed first and is then followed by space–time decoding. However, we will show that thanks to a more efficient linear transformation matrix $\mathbf{W}_i = \mathbf{D}_i^+ \mathbf{P}_i$, it can achieve a slight performance gain over the Type I detector. Throughout this letter, we shall refer to this detector as the Type I improved detector.

B. Type II Detector

In the Type II detector, we make use of the linear structure of the STBC to decode first. In particular, an equivalent channel is obtained after a series of linear transformations. The group detector is then applied to this equivalent channel.

Let $\mathbf{u}^{\mathrm{II}} = [\mathbf{y}_1', \mathbf{y}_2', \ldots, \mathbf{y}_T']'$, where $\mathbf{y}_i$ is the ith column vector of $\mathbf{Y}$, $i = 1, \ldots, T$. Then, from (2), we have

$$\begin{aligned} \mathbf{u}^{\mathrm{II}} = \sqrt{\frac{\mathrm{SNR}}{m}} \Bigg\{ &\begin{bmatrix} \mathbf{H}_1\mathbf{A}_{11} & \cdots & \mathbf{H}_G\mathbf{A}_{G1} \\ \vdots & \ddots & \vdots \\ \mathbf{H}_1\mathbf{A}_{1T} & \cdots & \mathbf{H}_G\mathbf{A}_{GT} \end{bmatrix} \cdot \begin{bmatrix} \mathbf{b}_1 \\ \vdots \\ \mathbf{b}_G \end{bmatrix} \\ &+ \begin{bmatrix} \mathbf{H}_1\mathbf{B}_{11} & \cdots & \mathbf{H}_G\mathbf{B}_{G1} \\ \vdots & \ddots & \vdots \\ \mathbf{H}_1\mathbf{B}_{1T} & \cdots & \mathbf{H}_G\mathbf{B}_{GT} \end{bmatrix} \cdot \begin{bmatrix} \mathbf{b}_1^* \\ \vdots \\ \mathbf{b}_G^* \end{bmatrix} \Bigg\} + \mathbf{v}^{\mathrm{II}}. \quad (17) \end{aligned}$$

Notice that $\{\mathbf{A}_{ij}\}_{j=1,\ldots,T_i}$ and $\{\mathbf{B}_{ij}\}_{j=1,\ldots,T_i}$ are the coefficient matrices of STBC_i, $i = 1, \ldots, G$. Assume that all the G groups have the same group size, and thus the same STBC.[2] Then, after a series of linear transformations, it can be shown that

$$\mathbf{r}^{\mathrm{II}} = \sqrt{\frac{\mathrm{SNR}}{m}} \left[\mathcal{H}_{\mathcal{G}_1}, \ldots, \mathcal{H}_{\mathcal{G}_G}\right] \cdot \mathbf{b} + \mathbf{z}^{\mathrm{II}} \quad (18)$$

where $\mathbf{r}^{\mathrm{II}} = \left[\left(\tilde{\mathbf{y}}_1^{\mathrm{II}}\right)', \ldots, \left(\tilde{\mathbf{y}}_T^{\mathrm{II}}\right)'\right]'$, and

$$\tilde{\mathbf{y}}_j^{\mathrm{II}} = \begin{cases} \mathbf{y}_j, & \text{if } \mathbf{B}_{ij} = \mathbf{0}_{m_i \times |\mathcal{G}_i|} \\ \mathbf{y}_j^*, & \text{if } \mathbf{A}_{ij} = \mathbf{0}_{m_i \times |\mathcal{G}_i|} \end{cases}, \quad j = 1, \ldots, T.$$

Also, $\mathbf{z}^{\mathrm{II}}$ has the covariance $\mathbf{I}_{Tn}$. For any group $\mathcal{G}_i$, $i = 1, \ldots, G$, its corresponding subchannel matrix $\mathcal{H}_{\mathcal{G}_i}$ is given by

$$\mathcal{H}_{\mathcal{G}_i} = \begin{bmatrix} \mathbf{H}_i\mathbf{A}_{i1} + \mathbf{H}_i^*\mathbf{B}_{i1}^* \\ \vdots \\ \mathbf{H}_i\mathbf{A}_{iT} + \mathbf{H}_i^*\mathbf{B}_{iT}^* \end{bmatrix}. \quad (19)$$

The process of converting the codeword matrix $\mathbf{X}$ to $\mathbf{b}$ can be regarded as a decoding process. Therefore, from (18), it can be seen that after a series of linear transformations, the decoding process has been done. Next, let us apply GZF to the equivalent channel given by (18). Let $\mathbf{P}_{\mathcal{G}_i} = \mathbf{I}_{Tn} - \mathcal{H}_{\bar{\mathcal{G}}_i}(\mathcal{H}_{\bar{\mathcal{G}}_i}^+ \mathcal{H}_{\bar{\mathcal{G}}_i})^{-1}\mathcal{H}_{\bar{\mathcal{G}}_i}^+$. Then, using the linear transformation $\mathbf{W}_i = \mathcal{H}_{\mathcal{G}_i}^+ \mathbf{P}_{\mathcal{G}_i}$ on $\mathbf{r}^{\mathrm{II}}$, we have

$$\hat{\mathbf{r}}^{\mathrm{II}} = \mathbf{W}_i \mathbf{r}^{\mathrm{II}} = \sqrt{\frac{\mathrm{SNR}}{m}} \mathbf{Q}_{\mathcal{G}_i}^{-1} \cdot \mathbf{b}_i + \hat{\mathbf{z}}^{\mathrm{II}} \quad (20)$$

[2]For the case of variable-rate STBC, the decoding will be much more complicated. The real and imaginary parts of the channel matrix and codes should be separated, and two equivalent channels are obtained. Group detection is then applied to these two equivalent channels to get the real and imaginary parts of the desired symbol, respectively. In this letter, we assume that all the groups adopt the same STBC so as to simplify the analysis. However, it should be pointed out that the Type II detector can deal with the case of variable-rate STBC. We omit the details due to limited space.

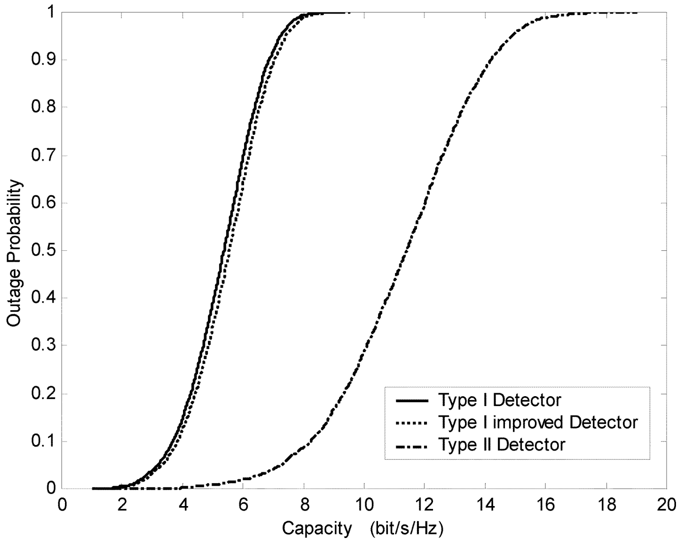

Fig. 2. Capacity cdf curves of Type I, Type I improved, and Type II detectors when $m = n = 4$, $G = 2$, and SNR $= 5$ dB.

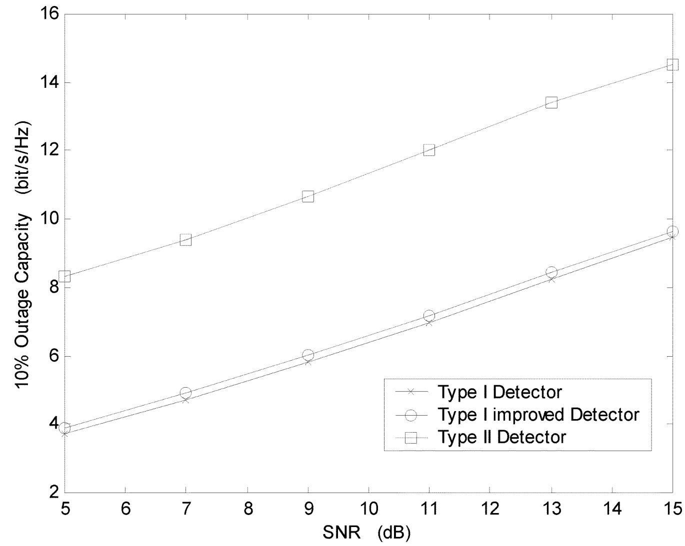

Fig. 3. 10% outage capacity versus SNR curves of Type I, Type I improved, and Type II detectors when $m = n = 4$ and $G = 2$.

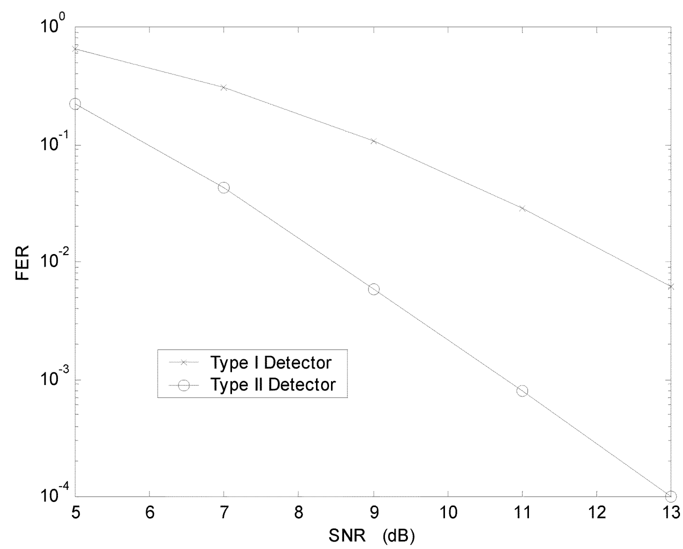

Fig. 4. FER versus SNR curves of Type I and Type II detectors when $m = n = 4$ and $G = 2$. QPSK is adopted.

where $\mathbf{Q}_{\mathcal{G}_i}^{-1} = \boldsymbol{\mathcal{H}}_{\mathcal{G}_i}^{+}\mathbf{P}_{\mathcal{G}_i}\boldsymbol{\mathcal{H}}_{\mathcal{G}_i}$ and $\mathrm{Cov}\left(\hat{\mathbf{z}}^{\mathrm{II}}\right) = \mathbf{Q}_{\mathcal{G}_i}^{-1}$. $\mathbf{b}_i$ can then be detected via a maximum-likelihood detector (MLD). The mutual information is therefore given by

$$\begin{aligned} I\left(\mathbf{b}_i;\hat{\mathbf{r}}^{\mathrm{II}}|\mathbf{H}{=}\mathbf{Q}_{\mathcal{G}_i}^{-1}\right) &= \log\det\left(\frac{\mathbf{Q}_{\mathcal{G}_i}^{-1}+\frac{\mathrm{SNR}}{m}\left(\mathbf{Q}_{\mathcal{G}_i}^{-1}\right)^{+}\mathbf{Q}_{\mathcal{G}_i}^{-1}}{\mathbf{Q}_{\mathcal{G}_i}^{-1}}\right)\\ &= \log\det\left(\mathbf{I}_{|\mathcal{G}_i|}+\frac{\mathrm{SNR}}{m}\left(\mathbf{Q}_{\mathcal{G}_i}^{-1}\right)^{+}\right). \quad (21)\end{aligned}$$

It can be seen that after decoding, the receive dimensions have increased to Tn, i.e., $\mathbf{r}^{\mathrm{II}}$ is $Tn \times 1$. Recall that in the Type I detector, group detection is applied to each $n \times 1$ receive signal vector $\mathbf{y}_i$, $i = 1, \ldots, T_i$. Obviously, in the Type II detector, better performance can be achieved by group detection, thanks to the higher receive dimensions. Considering that the entire performance is limited by the step of group detection,[3] it can be expected that the Type II detector can perform better than the Type I detector.

IV. Results and Discussions

The $x\%$ outage capacity $C_{\mathrm{out},x}$ is defined as the information rate that is guaranteed for $1-x\%$ of the channel realizations, i.e., $P(I \geq C_{\mathrm{out},x}) = 1 - x\%$. From (13), (16), and (21), we can compute the outage capacity achieved by the Type I detector, Type I improved detector, and Type II detector, respectively.

Assume that there are four transmit and four receive antennas, i.e., $m = 4$ and $n = 4$. The bit stream and transmit antennas are equally divided into $G = 2$ groups, and in each group, Alamouti's scheme is adopted. Fig. 2 presents the capacity cumulative distribution function (cdf) curves of the Type I, Type I improved, and Type II detectors for an SNR of 5 dB. It can be seen that compared with the Type I detector, the Type I improved detector can achieve a slight capacity gain. However, with the Type II detector, the capacity can be improved significantly. At least 4 b/s/Hz gain can be obtained. This is shown more clearly in Fig. 3, where 10% outage capacity against SNR curves are plotted. With an increasing SNR, more and more capacity gain can be achieved by the Type II detector over the Type I detector.

Fig. 4 further compares the FER performance of the Type I and II detectors, where quaternary phase-shift keying (QPSK) is assumed. It should be pointed out that in the Type I detector, the number of receive antennas n should be larger than $m - m_i$, because group detection is applied to an $n \times 1$ receive signal vector. However, for the Type II detector, it is only required that $Tn > m - m_i$, since group detection is performed after decoding. This implies that for some cases, when $n < m$, the Type I detector cannot work, while the Type II detector can. Here, for comparison, it is assumed that $n = m = 4$. As Fig. 4 shows, the Type II detector can achieve a gain of 4 dB over the Type I detector at an FER of 10^{-2}, and even more gains can

[3]The whole detection process can be divided into two steps: group detection and space–time block decoding. Group detection adopts ZF to suppress the intergroup interference. Compared with space–time block decoding (which achieves the same performance as MLD), group detection obviously has much worse performance. Therefore, we claim that "the entire performance is limited by the step of group detection."

TABLE I
COMPLEXITY COMPARISON OF TYPE I DETECTOR AND TYPE II DETECTOR

Detector	Complexity		
	Addition	Multiplication	Comparison
Type I	$\sum_{i=1}^{G} nm_i T_i + m_i^2 \lvert\mathcal{G}_i\rvert T_i + m_i \lvert\mathcal{G}_i\rvert T_i$	$\sum_{i=1}^{G} nm_i T_i + m_i^2 \lvert\mathcal{G}_i\rvert T_i + m_i \lvert\mathcal{G}_i\rvert T_i$	$K\Omega$ *
Type II	$\sum_{i=1}^{G} n\lvert\mathcal{G}_i\rvert T + nm_i \lvert\mathcal{G}_i\rvert T$	$\sum_{i=1}^{G} n\lvert\mathcal{G}_i\rvert T + nm_i \lvert\mathcal{G}_i\rvert T$	$\sum_{i=1}^{G} \Omega^{\lvert\mathcal{G}_i\rvert}$

* Ω is the constellation size. For example, $\Omega=4$ for QPSK modulation.

be achieved with increasing SNR, thanks to its higher diversity order. It should be pointed out that group successive interference cancellation (GSIC) can be also applied to both the Type I and Type II detectors. With GSIC, the performance of both detectors will be greatly improved. However, the gap between these two detectors is still large. For example, when $n = m = 4$, a 3-dB gain can be achieved by the Type II detector at an FER of 10^{-2}. We omit the results for GSIC here, due to limited space.

We conclude this section by noting that in the Type II detector, a $|\mathcal{G}_i|$-symbol MLD inside each group is necessary. The advantage of the STBC low-complexity detection cannot be exploited, and thus, the proposed Type II detector leads to a complexity increase, compared with the traditional one. In Table I, we list the number of additions, multiplications, and comparisons required by the Type I and Type II detectors, respectively. From Table I, it can be seen that the Type I and Type II detectors require approximately an equal number of additions and multiplications. However, the number of comparisons of the Type I detector is much less than that of the Type II detector, due to the use of linear detection for each group in the Type I detector. For example, with the above parameters, 80 and 96 multiplications and additions are needed by Type I and Type II detectors, respectively. However, the number of comparisons of the Type II detector is double of that of the Type I detector (16 and 32 comparisons for Type I and Type II detectors, respectively). Nevertheless, note that for a small group size, the complexity of the Type II detector is comparable to that of the Type I detector.

V. CONCLUSIONS

GLST is an architecture based on group transmission and group detection, where a good tradeoff between multiplexing gain and diversity gain can be achieved. In this letter, a novel detector for GLST was proposed, where decoding is performed before group detection. It is shown that compared with the traditional Type I detector for GLST, the proposed Type II detector can achieve significant capacity gains and much higher diversity order, due to its higher receive dimensions. In addition, the Type II detector has a lower requirement on the number of receive antennas. Such superior performance is achieved, however, at the expense of higher complexity.

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