

Max-Log-MAP Optimal MU-MIMO Receiver for Joint Data Detection and Interferer Modulation Classification

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Abstract—We consider a multi-user multiple-input multiple-output (MU-MIMO) system that uses orthogonal frequency division multiplexing (OFDM). Several receivers are developed for data detection, where the receivers have partial state information about the MU-MIMO transmission; however, the modulation constellation of the co-scheduled users is unknown. We propose a joint data detection and maximum likelihood (ML) modulation classification (MC) of the co-scheduled user. We show that the decision metric for MC is an accumulation of the Euclidean distances used by the ML single-user MIMO detector. An efficient hardware architecture emerges that exploits this commonality between the classification and detection steps and results in sharing of hardware resources.

Index Terms—Max-Log-MAP, MIMO detection.

I. INTRODUCTION

MU-MIMO was proposed as a method for increasing the capacity of wireless networks [1]. The LTE standard uses MIMO and OFDM access to schedule multiple users for transmission on the same physical resource blocks (PRBs) [2]. Users are paired together such that their cascaded precoder-channel vectors are maximally orthogonal. The full MU-MIMO channel can be estimated using pilots transmitted by the base station. The inability of the two users' signals to appear as orthogonal waveforms to the receiver stems from: 1) the finite set of available precoders specified by the standard (to satisfy limited information feedback) and 2) the delay between the time users are co-scheduled and the time the MU-MIMO transmission occurs, the channel state might change (typical in high Doppler scenarios).

Several receiver processing methods at the user equipment (UE) have been proposed for MU-MIMO systems. One approach is to treat the co-scheduled user in MU-MIMO just as additional noise. This type of processing results in sub-optimal receiver performance as discussed in [1]. Another approach [3], [4] is to employ linear processing using minimum mean-squared error (MMSE) or interference rejection combining (IRC), where only knowledge of the channel estimate of the co-scheduled interfering user is used to detect the desired user. In [5]–[7], the size of the interfering user's constellation is always assumed 16-QAM regardless of its actual size, and an ML detector is used to separate the two users. In [8], the constellation size of the co-scheduled user

is estimated before joint detection. However, layer separation has to be applied first using zero-forcing or MMSE, whose performances are severely deteriorated in highly correlated channels [9]. The detector of [10] detects both users assuming knowledge of the interfering modulation type, and does not perform joint detection and modulation classification as we do in this letter.

Contributions: We develop *optimal* detection schemes for MU-MIMO systems without making any assumptions about the precoder design at the transmitter. We employ joint ML classification and detection for estimating the modulation constellation size of the co-scheduled user's signal, followed by data detection of the desired user. The novelty is to treat the unknown co-scheduled user's modulation constellation size as part of the MU-MIMO detector. Furthermore, we show that the optimal MU-MIMO detector can be *efficiently* implemented using the ML single-user MIMO (SU-MIMO) detector with a slight increase in complexity by exploiting the commonality between the constellation size classification and data detection steps. We also compare the proposed ML MU-MIMO receiver to the linear receiver approaches available in the literature [4].

The rest of the letter is organized as follows. The system model is described in Section II. Linear receivers are discussed in Section III, followed by the nulling receiver in Section IV. Our proposed joint optimal constellation size estimation and detection scheme is presented in Section V, and an efficient MU-MIMO receiver hardware architecture is discussed in Section VI. Section VII presents simulation results, while Section VIII concludes the letter.

II. SYSTEM MODEL

Consider an OFDM-based MU-MIMO system where K users are co-scheduled on the same PRBs. Let N denote the number of tones in each PRB. Each UE is equipped with N_r receive antennas and receives a single spatial stream. The received frequency-domain complex signal $\mathbf{y}_i \in \mathcal{C}^{N_r \times 1}$ at the UE of interest on the i th resource element (RE) over which the K users are scheduled is given by $\mathbf{y}_i = \mathbf{H}_i \mathbf{x}_i + \mathbf{n}_i$, $i = 1, \dots, N$, where $\mathbf{H}_i \in \mathcal{C}^{N_r \times K}$ is the complex channel matrix, \mathbf{x}_i denotes the transmitted $K \times 1$ QAM symbol vector for K users, and $\mathbf{n}_i \in \mathcal{C}^{N_r \times 1}$ denotes thermal noise modeled as a zero-mean complex Gaussian random vector with covariance matrix $\mathbf{R} \triangleq \mathbb{E}[\mathbf{n}_i \mathbf{n}_i^*] = \sigma^2 \mathbf{I}_{N_r}$.

We consider the practical case where $K = N_r = 2$, thus $\mathbf{x}_i = [x_1^{(i)} \ x_2^{(i)}]^T$ and $\mathbf{y}_i = \mathbf{h}_1^{(i)} x_1^{(i)} + \mathbf{h}_2^{(i)} x_2^{(i)} + \mathbf{n}_i$, where $\mathbf{H}_i \triangleq [\mathbf{h}_1^{(i)} \ \mathbf{h}_2^{(i)}]$, and $\mathbf{h}_1^{(i)}$ and $\mathbf{h}_2^{(i)}$ denote the cascade of the channel and the precoders of User-1 (U1) and User-2 (U2), respectively. Let U1 denote the user of interest while U2 denote the interfering (co-scheduled) user whose constellation size is unknown to U1. Let M_S and M_I be the constellations

Manuscript received February 11, 2016; accepted April 6, 2016. Date of publication April 11, 2016; date of current version July 8, 2016. The associate editor coordinating the review of this letter and approving it for publication was T. Ngatched.

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Digital Object Identifier 10.1109/LCOMM.2016.2552514

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of U1 and U2, respectively, and N the number of REs over which M_I is constant. The symbols $x_1^{(i)}$ and $x_2^{(i)}$ represent $\log_2(|M_S|)$ and $\log_2(|M_I|)$ coded bits from a channel encoder (e.g., turbo code). The j th bit of symbol $x_1^{(i)}$ is denoted by b_{ij} .

III. LINEAR RECEIVER METHODS

Two linear MU-MIMO detection methods are reviewed. They differ in complexity and performance.

A. Covariance-Based Linear Receiver

This receiver estimates the interference covariance matrix $\mathbf{R}_{\mathbf{u}\mathbf{u}}$ using knowledge of only U1's channel, thus having the least knowledge about U2. Denote $x_1^{(p)}$ and \mathbf{y}_p as the known pilot and received vector, respectively, on the p th tone, where $p = 1, \dots, N_p$ and N_p is the number of pilot tones amongst the N tones over which the two users are co-scheduled. Then, the sample $\mathbf{R}_{\mathbf{u}\mathbf{u}}$ matrix is estimated as

$$\mathbf{R}_{\mathbf{u}\mathbf{u}} = \frac{1}{N_p} \sum_{p=1}^{N_p} \mathbf{u}_p \mathbf{u}_p^*,$$

where $\mathbf{u}_p = \mathbf{y}_p - \mathbf{h}_1^{(p)} x_1^{(p)}$. The weight vector for the linear receiver of U1 is given by $\mathbf{w}_{1,\text{cov}}^{(i)} = \mathbf{R}_{\mathbf{u}\mathbf{u}}^{-1} \mathbf{h}_1^{(i)}$ which is applied to \mathbf{y}_i to obtain $z_{1,\text{cov}} \triangleq \mathbf{w}_{1,\text{cov}}^{(i)*} \mathbf{y}_i$ that is used to generate LLR values.

B. Linear IRC Receiver

An improvement over the covariance-based linear receiver is obtained wherein the UE receiver uses estimates of *both* the desired channel as well as the co-scheduled user's channel. For covariance matrix \mathbf{R} , we arrive at an IRC weight vector that is given by [3], [4]

$$\mathbf{w}_{1,\text{IRC}}^{(i)} = \left(\mathbf{h}_2^{(i)} \mathbf{h}_2^{(i)*} + \mathbf{R} \right)^{-1} \mathbf{h}_1^{(i)}.$$

The desired user receiver processing is obtained by $z_{1,\text{IRC}}^{(i)} = \mathbf{w}_{1,\text{IRC}}^{(i)*} \mathbf{y}_i$, which can be approximated by a complex Gaussian random variable, i.e. $p\left(z_{1,\text{IRC}}^{(i)} | x_1^{(i)}, \mathbf{h}_1^{(i)}, \mathbf{h}_2^{(i)}\right) \sim \mathcal{CN}(\mu_z, v_z^2)$, with mean $\mu_z = \mathbf{w}_{1,\text{IRC}}^{(i)*} \mathbf{h}_1^{(i)} x_1^{(i)}$ and variance $v_z^2 = \mathbf{w}_{1,\text{IRC}}^{(i)*} \left(\mathbf{h}_2^{(i)} \mathbf{h}_2^{(i)*} + \mathbf{R} \right) \mathbf{w}_{1,\text{IRC}}^{(i)}$. The LLR of the j th bit of the desired user QAM symbol $x_1^{(i)}$ using the max-log-MAP approximation [11] is given by

$$\Lambda(b_{ij}) = \frac{1}{v_z^2} \left[\min_{x_1^{(i)} \in M_S^{(-1)}} |z_{1,\text{IRC}}^{(i)} - \mu_z|^2 - \min_{x_1^{(i)} \in M_S^{(+1)}} |z_{1,\text{IRC}}^{(i)} - \mu_z|^2 \right]$$

where $M_S^{(+1)} = \{x_1^{(i)} : b_{ij} = +1\}$ and $M_S^{(-1)} = \{x_1^{(i)} : b_{ij} = -1\}$.

IV. NULLING RECEIVER FOR CONSTELLATION SIZE ESTIMATION

The constellation size of U2 can be estimated by nulling out the *unknown* signal of U1 using a linear filter [8]. The equivalent system after applying this linear filter to the received vector is given by

$$\tilde{\mathbf{y}}_i = \mathbf{g}_i^* \mathbf{y}_i = \mathbf{g}_i^* \mathbf{h}_2^{(i)} x_2^{(i)} + \mathbf{g}_i^* \mathbf{n}_i \triangleq a_i x_2^{(i)} + \tilde{n}_i,$$

where $\mathbf{g}_i \in \mathcal{C}^{2 \times 1}$ is chosen to be orthogonal to $\mathbf{h}_1^{(i)}$, i.e., $\mathbf{g}_i^* \mathbf{h}_1^{(i)} = 0$. The nulling filter is given by $\mathbf{g}_i = \mathbf{G}\mathbf{c}$, where

\mathbf{c} is a 2×1 combining vector and \mathbf{G} is the left null-space projection matrix given by

$$\mathbf{G} = \mathbf{I}_2 - \mathbf{h}_1^{(i)} \mathbf{h}_1^{(i)*} / \|\mathbf{h}_1^{(i)}\|^2.$$

Since \mathbf{G} has rank 1, the combining vector can simply be $\mathbf{c} = [1 \ 0]^T$.

Following a similar derivation to Section V, the ML estimate of the co-scheduled user's constellation using $\tilde{\mathbf{y}}_i$ is given by

$$\hat{M}_I \approx \arg \min_{M_I \in \mathcal{M}} \left(\frac{N \log(|M_I|) + \sum_{i=1}^N \min_{x_2^{(i)} \in M_I} |\tilde{y}_i - a_i x_2^{(i)}|^2}{\mathbf{g}_i^* \mathbf{R} \mathbf{g}_i} \right).$$

Once the co-scheduled user constellation \hat{M}_I is estimated, the LLRs of the desired user are generated (see (2) in Section V).

V. PROPOSED JOINT ESTIMATION AND DETECTION SCHEME

The key idea in the proposed receiver is to better exploit the diversity offered by the two receive antennas. We develop an optimal receiver for user 1 through estimating the modulation constellation size of user 2 using the received signal \mathbf{y}_i and knowledge of the MIMO channel. The optimal ML estimator will inherently average out the unknown desired signal using knowledge of its constellation size. The new ML estimate of the interfering user constellation is the one whose joint probability with the observation vector $\{\mathbf{y}_i\}_{i=1}^N$ is maximum [12]. With no prior information about the likelihood of each constellation size, their apriori probabilities are equally set. Hence, maximizing the joint distribution boils down to maximizing the distribution of the observation signal conditioned on the unknown constellation size as follows:

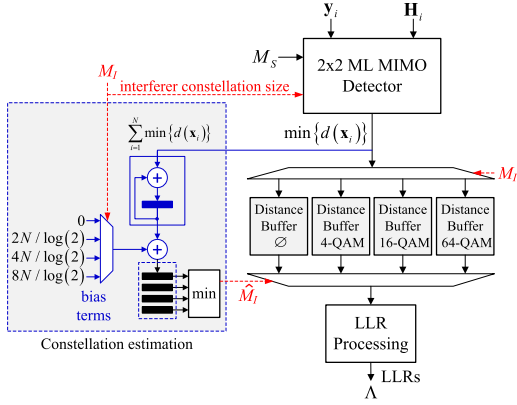
$$\hat{M}_I = \arg \max_{M_I \in \mathcal{M}} p\left(\{\mathbf{y}_i\}_{i=1}^N \mid \{\mathbf{H}_i\}_{i=1}^N, M_S, M_I\right),$$

where $\mathcal{M} \triangleq \{\emptyset, 4\text{-QAM}, 16\text{-QAM}, 64\text{-QAM}\}$ denotes the set of allowable constellations that the interferer can assume, including the case when the co-scheduled user is not present, i.e. \emptyset . Since $x_1^{(i)}, x_2^{(i)}$ and \mathbf{n}_i are independent for $i = 1, \dots, N$, the ML estimate of the interferer's constellation can then be written as

$$\begin{aligned} \hat{M}_I &= \arg \max_{M_I \in \mathcal{M}} \prod_{i=1}^N p(\mathbf{y}_i | \mathbf{H}_i, M_S, M_I) \\ &= \arg \max_{M_I \in \mathcal{M}} \prod_{i=1}^N \sum_{x_1^{(i)} \in M_S} \sum_{x_2^{(i)} \in M_I} p(\mathbf{y}_i | \mathbf{H}_i, M_S, M_I, x_1^{(i)}, x_2^{(i)}) p(x_1^{(i)}) p(x_2^{(i)}) \\ &= \arg \max_{M_I \in \mathcal{M}} \frac{1}{|M_I|^N} \prod_{i=1}^N \sum_{x_1^{(i)} \in M_S} \sum_{x_2^{(i)} \in M_I} p(\mathbf{y}_i | \mathbf{H}_i, M_S, M_I, x_1^{(i)}, x_2^{(i)}) \end{aligned}$$

where $|M_I|$ denotes the size of the interfering user constellation. The last equality follows from the assumptions that

$$p(x_1^{(i)}) = \frac{1}{|M_S|}, \quad \text{and} \quad p(x_2^{(i)}) = \frac{1}{|M_I|}, \quad \forall i.$$

Fig. 1. 2×2 ML MU-MIMO detector using a SU-MIMO detector as a core.

Let $\mathbf{x}_i = [x_1^{(i)} x_2^{(i)}]^T$ and $d(\mathbf{x}_i) = (\mathbf{y}_i - \mathbf{H}_i \mathbf{x}_i)^* \mathbf{R}^{-1} (\mathbf{y}_i - \mathbf{H}_i \mathbf{x}_i)$, we can then write \hat{M}_I as

$$\hat{M}_I = \arg \max_{M_I \in \mathcal{M}} \frac{1}{|M_I|^N} \prod_{i=1}^N \sum_{x_1^{(i)} \in M_S} \sum_{x_2^{(i)} \in M_I} \exp(-d(\mathbf{x}_i)).$$

Using the max-log-MAP approximation [11], we have that

$$\hat{M}_I \approx \arg \min_{M_I \in \mathcal{M}} \left(N \log(|M_I|) + \sum_{i=1}^N \min_{x_1^{(i)} \in M_S, x_2^{(i)} \in M_I} d(\mathbf{x}_i) \right), \quad (1)$$

where $\log(\cdot)$ is the natural logarithmic function.

As seen from (1), the decision metric to estimate the interferer's constellation is the sum of the minimum Euclidean distances $d(\mathbf{x}_i)$ between the received vector \mathbf{y}_i and the symbol vector \mathbf{x}_i for the set of N tones over which the co-scheduled user is stationary (fixed constellation). Once the co-scheduled user's constellation, \hat{M}_I , is estimated, then the LLR of the j th bit of the desired user QAM symbol $x_1^{(i)}$ using the max-log-MAP approximation is given by [13]

$$\Lambda(b_{ij}) = \min_{x_1^{(i)} \in M_S^{(-1)}, x_2^{(i)} \in \hat{M}_I} d(\mathbf{x}_i) - \min_{x_1^{(i)} \in M_S^{(+1)}, x_2^{(i)} \in \hat{M}_I} d(\mathbf{x}_i), \quad (2)$$

where $M_S^{(+1)}$ and $M_S^{(-1)}$ are defined in Section III-B. As seen from (2), computing the LLRs involves the same Euclidean distance computations as those needed for the co-scheduled user's constellation estimation in (1). This fact will be exploited in the hardware implementation of the proposed MU-MIMO receiver.

VI. EFFICIENT RECEIVER ARCHITECTURE

We present an optimized architecture for the proposed 2×2 MU-MIMO detector in Fig. 1, using a 2×2 ML SU-MIMO detector core. The 2×2 ML SU-MIMO detector is one that has knowledge of the modulation constellation size on both of its layers. The MU-MIMO detector processes the received vector \mathbf{y}_i assuming all $|\mathcal{M}| = 4$ possible choices of the interferer's constellation. It generates 4 corresponding lists of minimum distance metrics formed from the sets $d(\mathbf{x}_i)$ for all the $M_I \in \mathcal{M}$ possible hypotheses of interfering user constellation, with $x_1^{(i)} \in M_S$. These distances are stored in four buffers each of size $|M_S|$.

TABLE I
COMPLEXITY OF PROPOSED VS. BRUTE-FORCE APPROACH

| Metric | Scheme | Primary User Constellation | | |
|-----------|-------------|----------------------------|--------|--------|
| | | QPSK | 16-QAM | 64-QAM |
| Distances | Brute-force | 1856 | 7424 | 29696 |
| | Proposed | 608 | 2432 | 9728 |
| Buffers | Brute-force | 1856 | 7424 | 29696 |
| | Proposed | 192 | 768 | 3072 |

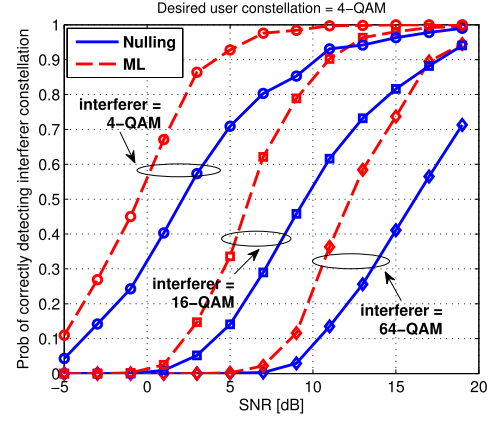


Fig. 2. Probability of correct interferer modulation constellation detection versus SNR. Desired user constellation is fixed to 4-QAM.

The minimum distance from each list is passed to an adder that accumulates the minimum distances over a span of N tones. The resulting 4 minimum accumulated distances for each interferer hypothesis are stored in a buffer. The minimum from this buffer is used to identify the interferer's constellation, and the corresponding stored distances in the buffers are selected and forwarded for LLR processing according to (2).

Since the interferer's MC remains static over N tones for a duration of 1 LTE subframe (14 OFDM symbols), the particular choice of $N = 12$ yields an accurate constellation estimation of U2 (see Section VII) and results in substantial savings in computations. Therefore, the detector only needs to run in the above mode to identify the interferer's constellation for one OFDM symbol in a subframe. Before switching back to normal ML SU-MIMO detection mode.

Taking the LTE scenario for hardware complexity analysis, the total number of possible tones in 1 PRB in a subframe is $12 \times 11 = 132$ assuming that 3 symbols are allocated for the control channel. Of these tones, 16 are reserved for pilots and 116 for data. In the architecture of Fig. 1, the total number of distance computations needed to generate LLRs from these data tones is $(116 - 12 + 4 \times 12) \times |M_S| = 152 \times |M_S|$. This corresponds to an increase of only 31% compared to the $116 \times |M_S|$ distances computed by a 2×2 ML SU-MIMO detector. A brute-force ML MU-MIMO detector that does not exploit our proposed sharing of hardware resources, would compute and buffer all $116 \times 4 \times |M_S|$ distances, which amounts to an increase in complexity by a factor of 3.1. Table I compares the complexity of the proposed approach versus the brute-force approach.

VII. SIMULATION RESULTS

We simulate the probability of correctly detecting the constellation size of the interfering user for different desired

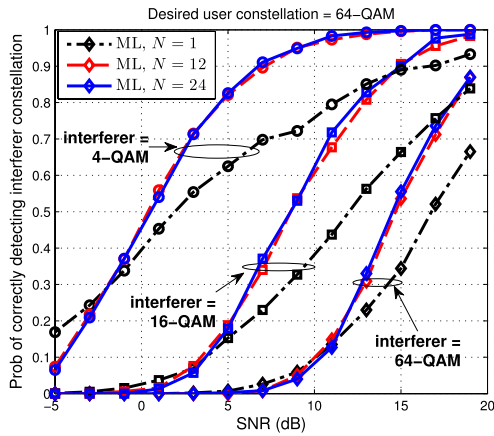


Fig. 3. Probability of correct interferer modulation constellation detection versus SNR (dB) for $N = 1, 12, 24$ tones and desired user is 64-QAM.

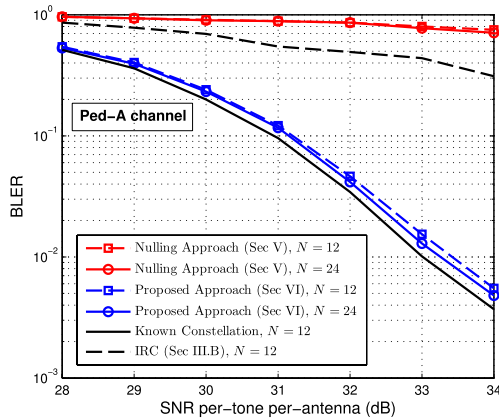


Fig. 4. BLER versus per-tone per-antenna SNR (dB). EPA channel with high correlation (0.9), 64-QAM for both users, and code-rate 1/2.

and interfering constellation sizes. The nulling receiver method is compared to the proposed method. The received desired user power is assumed to be equal to the interfering user's power, i.e. the signal-to-interference ratio (SIR) is 0dB. The channel is generated as a $\mathcal{CN}(0, 1)$ random variable and is i.i.d. from tone-to-tone. A random precoder is applied at the transmitter and the channel is assumed to be known at the receiver. Figure 2 shows the results for $M_S = 4$ -QAM with $M_I = 4, 16, 64$ -QAM using $N = 24$ REs. Our ML classification method has a 5 dB gain over the nulling approach for $M_S = 4$ -QAM.

In Fig. 3 we compare the constellation size detection performance with $N = 1, 12$ and 24 for $|M_S| = 64$ and different interference constellation sizes. Figure 3 shows that $N = 12$ REs are adequate for robust detection of the interferer's constellation size.

Remark: Increasing N should in general improve performance due to better noise averaging. However, in addition to noise, there is residual error due to the minimization of the quadratic term in (1). This error occurs when the \mathbf{x}_i that minimizes the quadratic term is different from the originally transmitted \mathbf{x} as a result of the noise exceeding the minimum distance of the lattice constructed

by the set of all symbol vectors. Notice that the detection performance with $N = 1$ is better than that with $N = 12$ and 24 for very low SNR due to the accumulation of noise rather than signal.

The BLER performance of the various receivers is compared when both users use 64-QAM, using a rate-1/2 LTE-turbo code. A random precoder is assumed for both users, and the channel is assumed to be known at the receiver. Figure 4 shows that our proposed ML MU-MIMO receiver is a fraction of dB away from having perfect knowledge of the co-scheduled user's constellation. It also has about several dBs gain over the nulling method that shows degradation over the MMSE method due to high channel correlation.

VIII. CONCLUSIONS

An OFDM receiver for MU-MIMO transmission where two users share the same time and frequency resources has been developed. This receiver is based on the joint ML modulation classification of the co-scheduled user and data detection, and outperforms linear receiver approaches. It has been shown that the decision metric for MC is an accumulation over a set of tones of minimum distance metrics, which are also used in the ML SU-MIMO detector for LLR computations. An efficient hardware architecture emerges that exploits this commonality between the MC and data detection steps, and results in sharing of hardware resources.

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