NEW ST-BC MIMO-CDMA TRANSCEIVER WITH AUGMENTED USER SIGNATURES

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ABSTRACT

This paper presents new transceiver framework for the DS-CDMA systems that use the multiple-input multiple-output (MIMO) antennas, with space-time block code (ST-BC). In the transmitter, new hybrid (prefix/postfix) zero-padding sequences combining with the desired user signature is exploited to form the hybrid augmented user signature sequences, with respect to different transmit-antennas. While in the receiver under the GSC framework a new Capon-like semi-blind MIMO-CDMA filterbank detector based on the linearly constrained constant modulus (LCCM) criterion, is derived. Also, the adaptive GSC-RLS algorithm is developed for adaptively implementing the blind LCCM MIMO-CDMA filterbank of the receiver. With the specific design of the hybrid augmented user signature sequences, we are able to resolve phase ambiguity problem of blind channel estimation. Computer simulations verify the merits of the proposed new transceiver framework; it can be employed to achieve better performance, and is more robust against the discrepancy due to Capon channel estimation.

1. INTRODUCTION

With the multiple transmit-antennas (*Tx*) and receive-antennas (*Rx*) (MIMO) technology, channel capacity of the direct sequence code division multiple access (DS-CDMA) systems can be increased, effectively, for wireless communications [1]-[4]. Also, the MIMO systems with space-time coding (STC) make spatial diversity possible to be exploited in downlink transmission since it relies on multiple transmit-antennas, which is feasible at the base-station [3]-[5]. The multiple access interference (MAI) is known to be the main performance limitation of the CDMA systems. In addition, the self-interference (SI) (due to ST-coding) induced by multiple transmit antennas in the ST-BC framework is another important issue. The blind receiver is an alternative candidate employed to increase spectral efficiency, since the training sequence is not required. However, the conventional blind receiver has inherent phase ambiguity problem [6][7]. To solve this problem, in this paper, new transceiver framework of the ST-BC MIMO-CDMA system is proposed. With specific design of new hybrid (prefix/postfix) zero-padding sequences combining with the desired user signatures in the transmitter, it can be used to alleviate the effect of MAI and SI. Also, associated with a new Capon-like blind channel estimator in the receiver, it can solve the phase ambiguity problem. Next, under the generalized sidelobe canceller (GSC) structure the GSC-RLS algorithm is derived for implementing the blind MIMO-CDMA filterbank of the receiver, adaptively.

2. NEW TRANSCEIVER FOR ST-BC MIMO-CDMA SYSTEM WITH LCCM GSC-RLS ALGORITHM

2.1 Model Description of ST-BC MIMO-CDMA System

Let us consider a synchronous downlink K-user ST-BC MIMO-CDMA system with two transmit-antennas and N receive-antennas as depicted in Fig.1. Without loss of generality, the first user is assumed to be the desired one. Before transmission, the consecutive symbols $b_k(2t\text{-}1)$ and $b_k(2t)$ for k^{th} user are spreading by the signature code-sequences, $\mathbf{c}_{mk}$ (m =1 and 2) with length J with respect to two transmit-antennas. New hybrid (prefix/postfix) zero-padding sequences combined with the user signatures $\mathbf{c}_{mk}$ are exploited to alleviate the effects of ISI and SI in multipath channels. The transmitted signal vectors via the first antenna, which contain the messages of K users at block time t, can be expressed, respectively, as

$$\mathbf{s}_1(2t-1)=\sum_{k=1}^{K}\sqrt{\rho_k}\,b_k(2t-1)\begin{bmatrix}\mathbf{c}_{1k}\\ \mathbf{0}_{D\times 1}\end{bmatrix} \tag{1a}$$

and

$$\mathbf{s}_1(2t)=\sum_{k=1}^{K}\sqrt{\rho_k}\left(-b_k^*(2t)\begin{bmatrix}\mathbf{0}_{D\times 1}\\ \mathbf{c}_{1k}\end{bmatrix}+\begin{bmatrix}1\\ \mathbf{0}_{(J+D-1)\times 1}\end{bmatrix}\right) \tag{1b}$$

In (1) $\mathbf{s}_1(2t\text{-}1)$ is with dimension $(J+D)\times 1$, in which the redundant chip-sequence using zero-padding with block length D is inserted after spreading code sequence, $\mathbf{c}_{1k}$, of user k to form the augmented signature vector with dimension $J\times 1$. While for $\mathbf{s}_1(2t)$, due to the ST-coding, the zero-padding with same block length is inserted before spreading code sequence to form another augmented signature vector, plus an extra vector with the first element being unity and zeros, otherwise. The second term in the bracket on the right-side of (1b) is used for partial channel estimation to resolve the phase-ambiguity. Similarly, in the second transmit-antenna ($Tx2$), we simply insert zero-padding with block length D after/before the spreading code sequence $\mathbf{c}_{2k}$, of user k to form the augmented signature vectors related to consecutive symbols, and are denoted as

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$$\mathbf{s}_2(2t-1)=\sum_{k=1}^{K}\sqrt{\rho_k}\,b_k(2t)\begin{bmatrix}\mathbf{c}_{2k}\\ \hline \mathbf{0}_{D\times 1}\end{bmatrix}_{(J+D)\times 1} \tag{2a}$$

and

$$\mathbf{s}_2(2t)=\sum_{k=1}^{K}\sqrt{\rho_k}\,b_k^*(2t-1)\begin{bmatrix}\mathbf{0}_{D\times 1}\\ \hline \mathbf{c}_{2k}\end{bmatrix}_{(J+D)\times 1} \tag{2b}$$

where ρ_k denotes the average power of the kth user and assumed that D is greater or equal to L (channel order). Since the augmented signature vectors with respect to different transmit antennas of the ST-BC MIMO-CDMA system are with different format, thus named as the hybrid augmented user signature scheme.

For simplicity, we assume that the receiver is synchronized to the first path of desired user, and the delay corresponding to the lth path is denoted as lT_c where the chip time of spreading code is T_c. If the channel response between the mth transmit-antenna and nth receive-antenna, in a vector form, is denoted by $\mathbf{h}_{mn}$ with dimension $L\times 1$, the composite channel vector is defined as $\mathbf{r}_{kmn}=[r_{kmn,1},\ \ldots,\ r_{kmn,q}]^T=\mathbf{c}_{mk} * \mathbf{h}_{mn} = \mathbf{C}_{mk}\mathbf{h}_{mn}$, where the $q\times L$ code matrix $\mathbf{C}_{mk}$ is a circular matrix, with its first column denoting as $[c_{mk1}\ldots c_{mkj}\ 0\ldots 0]^T$ and $q=J+L-1$. $\mathbf{P}_{mn}(2t)=\mathbf{s}_m(2t)*\mathbf{h}_{mn}$ and $\mathbf{P}_{mn}(2t-1)=\mathbf{s}_m(2t-1)*\mathbf{h}_{mn}$ are the convolution of signal vectors, $\mathbf{s}_m(2t-1)$ and $\mathbf{s}_m(2t)$ with the corresponding channel response $\mathbf{h}_{mn}$, respectively. For further discussion, we define two augmented interference vectors related to the previous and next interfering symbols, respectively,

$$\mathbf{r}_{kmn}^{(L)}=[r_{kmn,J+1},\cdots,r_{kmn,q},\underbrace{0,\cdots,0}_{J}]_{q\times 1}^T \tag{3a}$$

$$\mathbf{r}_{kmn}^{(R)}=[\underbrace{0,\cdots,0}_{J},r_{kmn,1},\cdots,r_{kmn,L-1}]_{q\times 1}^T \tag{3b}$$

They represent the effects of ISI. At the $2t$-1 time slot, after ST-BC encoder, the received signals vector at nth receive-antenna is given by

$$\mathbf{x}_n(2t-1)=\sum_{k=1}^{K}\left\{\sqrt{\rho_k}b_k(2t-1)\begin{bmatrix}\mathbf{r}_{k1n}\\ \hline \mathbf{0}_{D\times 1}\end{bmatrix}-\sqrt{\rho_k}b_k^*(2t-2)\begin{bmatrix}\mathbf{r}_{k1n}^{(L)}\\ \hline \mathbf{0}_{D\times 1}\end{bmatrix}-\sqrt{\rho_k}\begin{bmatrix}\mathbf{0}_{(J+D)\times 1}\\ \hline h_{1n,1}\\ \vdots\\ h_{1n,L-1}\end{bmatrix}\right. \tag{4}$$
$$\left.+\sqrt{\rho_k}b_k(2t)\begin{bmatrix}\mathbf{r}_{k2n}\\ \hline \mathbf{0}_{D\times 1}\end{bmatrix}+\sqrt{\rho_k}b_k^*(2t-3)\begin{bmatrix}\mathbf{r}_{k2n}^{(L)}\\ \hline \mathbf{0}_{D\times 1}\end{bmatrix}\right\}+\mathbf{w}_n(2t-1)$$

and

$$\mathbf{x}_n(2t)=\sum_{k=1}^{K}\sqrt{\rho_k}\left\{-b_k^*(2t-1)\begin{bmatrix}\mathbf{0}_{D\times 1}\\ \hline \mathbf{r}_{k1n}\end{bmatrix}+\begin{bmatrix}h_{1n,1}\\ \vdots\\ h_{1n,L}\\ \hline \mathbf{0}_{(J+D-1)\times 1}\end{bmatrix}+b_k(2t+1)\begin{bmatrix}\mathbf{0}_{D\times 1}\\ \hline \mathbf{r}_{k1n}^{(R)}\end{bmatrix}\right. \tag{5}$$
$$\left.+b_k^*(2t-1)\begin{bmatrix}\mathbf{0}_{D\times 1}\\ \hline \mathbf{r}_{k2n}\end{bmatrix}_{(q+D)\times 1}+b_k(2t+2)\begin{bmatrix}\mathbf{0}_{D\times 1}\\ \hline \mathbf{r}_{k2n}^{(R)}\end{bmatrix}\right\}+\mathbf{w}_n(2t)$$

Now, we would like to exploit how to partially remove the effects of ISI and self-interference. First, in (4) five terms are included in the bracket, and except the third term (it contains the information of channel coefficients) and the noise vector, the last D samples of $\mathbf{x}_n(2t\text{-}1)$ are nulls. Similarly, in (5) except the second term inside the bracket and noise vector, the first D samples of all other terms contained in $\mathbf{x}_n(2t)$ are nulls. Thus, by removing the redundancy from the received signal vector, i.e., the last/first D samples of the received signal vector $\mathbf{x}_n(2t\text{-}1)$ / $\mathbf{x}_n(2t)$, we can enhance the system performance by depressing the effects of MAI, ISI and self-interference. In consequent, we obtain the reduced received signal vectors at Rxn:

$$\mathbf{x}'_n(2t-1)=\sum_{k=1}^{K}\sqrt{\rho_k}\left\{b_k(2t-1)\mathbf{r}_{k1n}-b_k^*(2t-2)\mathbf{r}_{k1n}^{(L)}\right. \tag{6}$$
$$\left.+b_k(2t)\mathbf{r}_{k2n}+b_k^*(2t-3)\mathbf{r}_{k2n}^{(L)}\right\}+\mathbf{w}'_n(2t-1)$$

and

$$\mathbf{x}'_n(2t)=\sum_{k=1}^{K}\left\{-\sqrt{\rho_k}b_k^*(2t)\mathbf{r}_{k1n}+\sqrt{\rho_k}b_k(2t+1)\mathbf{r}_{k1n}^{(R)}\right. \tag{7}$$
$$\left.+\sqrt{\rho_k}b_k^*(2t-1)\mathbf{r}_{k2n}+\sqrt{\rho_k}b_k(2t+2)\mathbf{r}_{k2n}^{(R)}\right\}+\mathbf{w}'_n(2t)$$

By stacking the received signal vectors of (6) and (7), we form a new $2q$ x 1 received vector $\mathbf{y}_n(t)$, with respect to the desired symbols $b_k(2t\text{-}1)$ and $b_k(2t)$ i.e.,

$$\mathbf{y}_n(t)=\begin{bmatrix}\mathbf{x}'_n(2t-1)\\ \mathbf{x}'^*_n(2t)\end{bmatrix}_{2q\times 1}$$
$$=\sum_{k=1}^{K}\sqrt{\rho_k}\left[b_k(2t-1)\mathbf{g}_{nk}+b_k(2t)\bar{\mathbf{g}}_{nk}\right]+\text{ISI}+\mathbf{v}_n(t) \tag{8}$$
$$=\sqrt{\rho_1}\left[b_1(2t-1)\mathbf{g}_{n1}+b_1(2t)\bar{\mathbf{g}}_{n1}\right]+\text{ISI}+\text{MAI}+\mathbf{v}_n(t)$$

where $\mathbf{v}_n=[\mathbf{w}_n^T(2t\text{-}1),\ \mathbf{w}_n^H(2t)]^T$ is the corresponding noise vector. Similarly, we may define the following parameters:

$\mathbf{h}_n=\left[\mathbf{h}_{1n}^T,\mathbf{h}_{2n}^H\right]^T$, $\mathbf{g}_{nk}=\left[\mathbf{r}_{k1n}^T,\mathbf{r}_{k2n}^H\right]^T=\boldsymbol{\Delta}_k\mathbf{h}_n$, $\boldsymbol{\Delta}_k=\begin{bmatrix}\mathbf{C}_{1k} & \mathbf{0}\\ \mathbf{0} & \mathbf{C}_{2k}\end{bmatrix}$,

$\bar{\mathbf{g}}_{nk}=\left[\mathbf{r}_{k2n}^T,-\mathbf{r}_{k1n}^H\right]^T=\bar{\boldsymbol{\Delta}}_k\mathbf{h}_n^*$, and $\bar{\boldsymbol{\Delta}}_k=\begin{bmatrix}\mathbf{0} & \mathbf{C}_{2k}\\ -\mathbf{C}_{1k} & \mathbf{0}\end{bmatrix}$.

By stacking the outputs of all receive-antennas, we obtain

$$\mathbf{y}(t)=\left[\mathbf{y}_1^T(t),\mathbf{y}_2^T(t),\cdots,\mathbf{y}_N^T(t)\right]^T$$
$$=\sqrt{\rho_1}\left[b_1(2t-1)\mathbf{g}_1+b_1(2t)\bar{\mathbf{g}}_1\right]+\boldsymbol{\gamma}(t) \tag{9}$$

This is the input to the multi-user detector. We note that the terms in the bracket on the right-hand side of (9) are the desired signals, and $\gamma(t)$ consists of the terms, MAI, ISI, and the stacked noise vector $\mathbf{v}(t)$. Similarly, the stacked vectors $\mathbf{g}_k(t)=\mathbf{D}_k\mathbf{h}(t)$, $\bar{\mathbf{g}}_k(t)=\bar{\mathbf{D}}_k\mathbf{h}^*(t)$, and $\mathbf{h}(t)$, are defined, accordingly, where $\mathbf{D}_k=\mathbf{I}_N\otimes\boldsymbol{\Delta}_k$ and $\bar{\mathbf{D}}_k=\mathbf{I}_N\otimes\bar{\boldsymbol{\Delta}}_k$. Now, based on the above formulation, following the approach as in [8][9] the problem becomes to design a two-branch linear filter $\mathbf{F}=\left[\mathbf{f}(t),\bar{\mathbf{f}}(t)\right]$, with dimension of $2q\times 1$, operating on the received vector $\mathbf{y}(t)$ to yield $\mathbf{z}_1(t)=\mathbf{F}^H\mathbf{y}(t)$, an estimate of $\mathbf{b}_1(t)=[b_1(2t\text{-}1),\ b_1(2t)]^T$. Both weight vectors $\mathbf{f}(t)$ and $\bar{\mathbf{f}}(t)$ of the filterbank are obtained by minimizing the filtered output power, subject to the unit-gain constraints with respect to the code matrices of the desired user, followed by the blind Capon channel estimator associated with our proposed hybrid redundant chip-sequences

2.2 MIMO CM-GSC-RLS Algorithm with Capon Channel Estimation

As in [8][9], in our design two branches of linear filterbank can operate, separately, thus we can compute tap weights of the two branches, independently. In what follows, only the derivation of the first branch is given, since with the similar approach the results can be inferred for the second branch. Next, with CM approach, output of the multi-user detector is assumed to be with constant envelop. The cost functions in the weighted least square (LS) form is defined as

$$J_{CM1}(\mathbf{f}(t)) = \sum_{i=1}^{t} \lambda^{t-i} \left| \alpha - \left| \mathbf{f}^H(t)\mathbf{y}(i) \right|^2 \right|^2 \quad (10)$$

where λ is the forgetting factor and t is number of iteration. The tap weights of $\mathbf{f}(t)$ in the receiver are obtained by minimizing (10) subjects to the constrain, $\mathbf{f}^H(t)\hat{\mathbf{g}}_1(t) = 1$, where $\hat{\mathbf{g}}_1(t)$ is the estimates of $\mathbf{g}_1(t)$ at iteration t. By substituting $\hat{\mathbf{g}}_1(t) = \mathbf{D}_1\hat{\mathbf{h}}(t)$ back into the constraint, and after some mathematical manipulation we have $\hat{\mathbf{h}}^H(t)\mathbf{D}_1^H\mathbf{f}(t) = 1$, where an estimate of $\mathbf{h}(t)$ is $\hat{\mathbf{h}}(t)$. If $\hat{\mathbf{h}}(t)$ is with unit-norm, the constraint becomes $\mathbf{D}_1^H\mathbf{f}(t) = \hat{\mathbf{h}}(t)$, where $\mathbf{D}_1$ is the code-constraint matrix. Eq.(10) can be reformulated as

$$\text{Minimize } J_{CM1}(\mathbf{f}(t)) \text{ subject to } \mathbf{D}_1^H\mathbf{f}(t) = \hat{\mathbf{h}}(t) \quad (11)$$

Next, we adopt the GSC structure to achieve better numerical stability and lower the computational complexity. Under the GSC structure the adaptive CM-GSC-RLS algorithm for ST-BC MIMO-CDMA receiver is derived, and the original tap weights, which satisfy the constraint, can be decomposed into $\mathbf{f}(t)=\mathbf{f}_c(t)-\mathbf{B}\mathbf{f}_a(t)$ [8][10], where $\mathbf{f}_c(t) = \mathbf{D}_1(\mathbf{D}_1^H\mathbf{D}_1)^{-1}\hat{\mathbf{h}}(t)$ is part which satisfies the constraint and is independent of data. The remaining part $-\mathbf{B}\mathbf{f}_a(t)$, which is uncorrelated with the constraint, can be viewed as our degree of freedom and can be adapted to suppress the overall interference $\boldsymbol{\gamma}(t)$. The blocking matrix $\mathbf{B}$ is chosen such that $\mathbf{D}_1^H\mathbf{B}=\mathbf{0}$. With above concepts in mind, for the first branch of the filterbank, the unconstrained optimization problem now becomes to minimize the cost function with respect to $\mathbf{f}_a(t)$. The overall of the optimal receiver can be illustrated as in Fig.2. By expanding $J_{CM1}(\mathbf{f}(t))$, we have

$$J_{CM1}(\mathbf{f}(t)) = \sum_{i=1}^{t}\lambda^{t-i}\left|\alpha - \left|\mathbf{f}^H(t)\mathbf{y}(i)\right|^2\right|^2 \cong \sum_{i=1}^{t}\lambda^{t-i}\left|\alpha - \mathbf{f}^H(t)\tilde{\mathbf{y}}(i)\right|^2 \quad (12)$$

where $\mathbf{f}^H(t)\mathbf{y}(t)$ is the output of first branch of the adaptive filterbank. The intermediate vector for the first branch is defined as $\tilde{\mathbf{y}}(t)=\mathbf{y}(t)\mathbf{y}^H(t)\mathbf{f}(t)$. The optimal solution of $\mathbf{f}_a(t)$ can be solved, by minimizing (12), with respect to $\mathbf{f}_a(t)$:

$$\mathbf{f}_{a,CM}(t) = \mathbf{R}_B^{-1}(t)\mathbf{B}^H(\mathbf{R}(t)\mathbf{f}_{c,CM}(t) - \boldsymbol{\theta}(t)) \quad (13)$$

The correlation matrix and cross-correlation vector are defined as

$$\mathbf{R}(t) = \lambda\mathbf{R}(t-1) + \tilde{\mathbf{y}}(t)\tilde{\mathbf{y}}^H(t) \quad \text{and} \quad \boldsymbol{\theta}(t) = \lambda\boldsymbol{\theta}(t-1) + \alpha\tilde{\mathbf{y}}(t) \quad ,$$

respectively, similarly, we define $\mathbf{f}_{c,CM}(t)=\mathbf{D}_1(\mathbf{D}_1^H\mathbf{D}_1)^{-1}\hat{\mathbf{h}}(t)$ and $\mathbf{R}_B(t)=\boldsymbol{B}^H\mathbf{R}(t)\boldsymbol{B}$. After mathematical manipulation, we obtain the recursive form of $\mathbf{f}_{a,CM}(t)$:

$$\mathbf{f}_{a,CM}(t) = \mathbf{f}_{a,CM}(t-1) - \mathbf{k}_B(t)e^*(t\,|\,t-1) + \boldsymbol{\Gamma}(t)\mathbf{f}_\varepsilon(t) \quad (14)$$

The LCCM GSC-RLS algorithm described above has the similar procedure as in Table 1 of [8]. Finally, we can use $\mathbf{f}_{CM}(t)$ = $\mathbf{f}_{c,CM}(t)$ - $\mathbf{B}$ $\mathbf{f}_{a,CM}(t)$ to obtain the final optimal tap weight. We note that the Capon channel estimation is accomplished by maximizing the filter output power after suppressing the overall interference. Similar results can be inferred for the second branch. Note that in Table I of [8] we derived a simple gradient scheme for tracking the value of α to improve the performance of the LCCM approach.

3. SIMULATION RESULTS

We consider a 10-user downlink CDMA system, with two transmit-antennas and N = 1 or 2 receive-antennas. The QPSK modulation and Gold code with length 31 and unit-energy are is considered. The noise coefficients are with zero-mean and the variances N_0 are determined according to SNR defined as E_s/N_0. Near-far ratio (NFR) is set to be 10 dB; it caused severe MAI, and the forgetting factor λ is set to be 0.998. The MMSE detection scheme with perfect (or known) channel information is the optimal solution and is used as the benchmark for comparison. To be consistent with [9], the true channel vector is employed for scaling $\hat{\mathbf{h}}(t)$, such that the first initial element of channel vector is error-free, that is the inherent ambiguity problem was ignored. For comparison, we also derived the so-called MV-GSC-RLS algorithm under new transceiver model for the ST-BC MIMO CDMA receiver, for adaptively implementing the LCMV-based blind Capon-type receiver. For fair comparison, the bit-error-rate (BER) of the proposed algorithm and MV-GSC-RLS algorithm [10] are computed by ignoring the first 500 iterations (corresponding to the 500 samples used in estimating the $\mathbf{R}_{yy}(t)$ for the blind Capon receiver) and ensured the proposed algorithm to converge.

First, we consider the case without mismatch, for L=3 as evident in Fig.3 the proposed LCCM-GSC-RLS algorithm could achieve better BER performance over the conventional Capon approach and MC-GSC-RLS algorithm. It has about 4dB improvements against the blind Capon receiver, and is about 2dB worst than the MMSE receiver (the benchmark) [12]. Next, the mismatch effect is investigated with similar parameters as in without mismatch case; the results are shown in Fig.4. In fact, the BER performances of the blind Capon receiver and MV-GSC-RLS algorithm are unsatisfactory under mismatch effect. This is true, when the value of SNR increases, the BER becomes worse than the proposed algorithm.

4. CONCLUSIONS

In this paper, we proposed new transceiver for the ST-BC MIMO-CDMA, where a specific hybrid augmented signature code-sequences of desired user was proposed, at transmitter. While in the receiver, a Capon-like adaptive blind MIMO-CDMA detector with two-branch filterbank, implemented by the proposed GSC-CM-RLS algorithm, was derived to resolve the phase ambiguity and combat the effects of ISI and self-interference. As demonstrated in computer simulation results, with this new transceiver framework we could achieve performance improvement and at the same time to solve the phase ambiguity problem, associated with blind channel estimation. It outperformed the conventional Capon receiver proposed in [9], with known true channel information, and the one with the MV-GSC-RLS (with the LCMV criterion), in terms of SINR and BER. This is especially true when the mismatch problem was considered in the MIMO-CDMA systems.

5. ACKNOWLEDGEMENT

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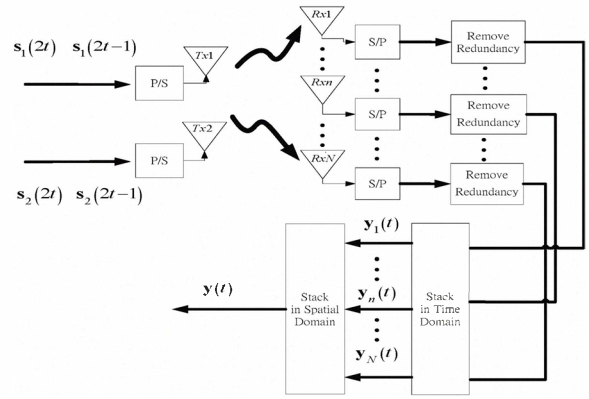

Figure 1 Configuration of the ST-BC MIMO-CDMA systems.

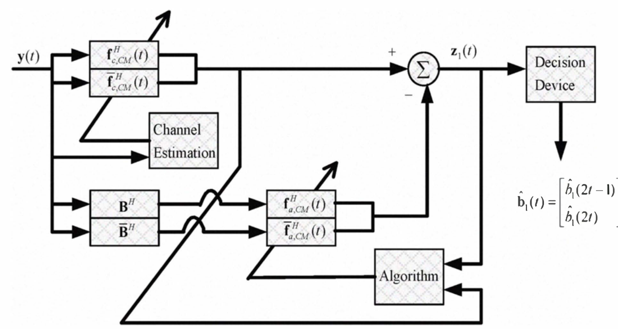

Figure 2 Configuration of the MIMO CM-GSC-RLS algorithm.

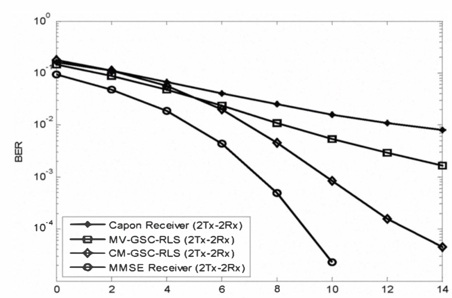

Figure 3 BER comparisons of different receivers without mismatch.

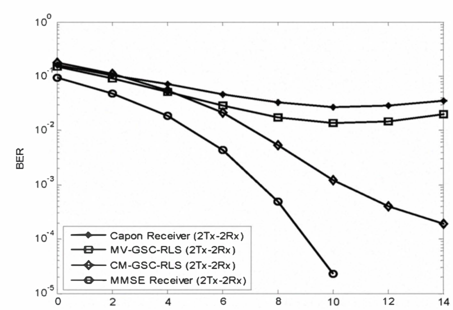

Figure 4 BER comparisons of different receivers with mismatch.