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寬頻無線通訊系統的錯誤控制機制之設計(1/3)

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主持人：蘇育德 交通大學教授 參與人員：吳建中、賴俊佑

Abstract--This paper proposes decoder structures for ST block codes. The decoders use the turbo principle to decode the "equivalent" serially- concatenated ST codes, assuming a frequency selective fading environment. We consider the more practical case issue when channel state has to be estimated. To begin with, we discuss the scenario when pilot symbols are available. Pseudo maximum likelihood (PML) and block least mean square (BLMS) methods are used to derive proper solutions. In order to increase bandwidth efficiency, i.e., reduce the length of transmitted pilot symbols, Viterbi algorithm (VA) will be associated with them for channel estimation. The second scenario is the blind case where there are no pilot symbols. We make use of the concept of tree search together with a branch and bound algorithm to reduce the complexity and make the overall algorithm practical. We show that a multiple-antenna system does deliver diversity gain it promises and iterative decoding procedure of (equivalent) serially-concatenated codes results in performance improvement.

I. INTORDUCTION

Mitigating the multi-path fading effect is the most challenging issue in designing a wireless communication receiver, as this effect results in unpredictable and time-varying amplitude attenuation and phase shift of the transmitted signal and introduces intersymbol interference (ISI). One of the simplest and perhaps, the most effective technique to enhance the receiver performance is the diversity scheme, which is usually categorized as temporal, frequency, and antenna (spatial) diversities.

It is well known that a good trade-off between coding gain and complexity can be achieved by serially concatenated codes proposed by Forney [1]. A serially concatenated code is one that applies two level of coding, an inner code and an outer code linked by an interleaver. The decoding principle of serially concatenated codes has been extended to joint equalization and decoding in an iterative form [2], [3] where the inner code is replaced by the equivalent convoluted channel. This method is referred to as "Turbo equalization". Just like in serially concatenated decoding, the improvement depends on the transmission of extrinsic information in an iterative process.

As a mobile communication system often operates in a time-varying environment, an adaptive receiver is required for channel estimation and tracking. Adaptive MLSE receiver needs perform joint channel estimation and data detection. An approach is proposed based on the MLSE using a block of N symbols at a time. The channel is updated once at the beginning of each block based on the survivor path from the previous block. This technique will be referred as block sequence estimation (BSE) [4]. Later, a new algorithm (pseudo maximum likelihood [PML]) based on the ML criterion that does not utilize the VA is proposed [5]. The PML algorithm is a symbol-by-symbol estimation algorithm within some delay and the channel state information CSI estimation is a part of the algorithm.

Almost all the blind schemes need complicated mathematical computation or copious operation formulation. Therefore exhaustive search method combined with branch and bound algorithm is proposed [6], [7]. This method solves a model by breaking up its possible region into successively smaller subsets (branching) and computing bounds on the objective value over each corresponding subset (bounding), and using them to remove some of the subsets from further consideration. In this paper we shall propose a scheme that can reduce search number to a reasonable value and find the ML solution the same.

The rest of this paper is organized as follows. Section II discusses several options to obtain channel estimate in a ST-coded systems. Section III discusses blind channel estimation. Numerical examples are presented to demonstrate the effectiveness of the proposed channel estimators. The last concludes this paper with a summary of our major contributions and suggestions for further studies.

II. Channel Estimation For Space-Time Coded System

Methods of channel estimation are often divided into two categories, i.e., schemes with training information and schemes without. The former estimators usually have faster convergence rate but less bandwidth efficiency and requires larger power consumption. Blind schemes have better bandwidth and power efficiency but need more computation and yield slower convergence.

A. Pilot-assisted channel estimators

We consider channel estimators with training information first. These estimators are also known as pilot-assisted channel estimators. Our proposed schemes utilize the concepts of maximum likelihood (ML) and LMS in conjunction with VA so that fewer training symbols are needed.

A.1 *Pseudo-maximum-likelihood (PML) algorithm*

A PML data estimation algorithm for discrete channels with finite memory is proposed in [5]. Unlike the traditional methods that utilize the VA for data sequence estimation, the PML algorithm offers an alternative solution to the problem. The PML algorithm is a symbol-by-symbol estimation algorithm. We will do some change for the original PML to adapt our proposal.

In the following equation r_j^k represents the signal of the *j*th receiver antenna at time *k* when a ST-coded waveform is transmitted over a multi-path fading channel with channel length *L*, N_t and N_r being the numbers of transmit and receive antennas, respectively.

(1) $$r_j^k = \sum_{i=1}^{N_t}\sum_{d=0}^{L-1} h_{i,j}^d a_i^{k-d} + n_j^k \quad j = 1,2,\cdots N_r.$$

The received samples at all receiver antennas can be expressed in a more compact

(2) $$\underline{r} = A\underline{h} + \underline{n}$$

where $\underline{r}$ is defined by

(3) $$\underline{r} = \left[\underline{r}_1^t \quad \underline{r}_2^t \quad \cdots \quad \underline{r}_{Nr}^t\right]^t$$

(4) $$\underline{r}_j = \left[r_j^{k+1} \quad r_j^{k+2} \quad \cdots \quad r_j^{k+N}\right]^t,$$

j=1,2,…,Nr

t denotes the transpose operation and *N* is the number of symbols per frame and it is assumed that *N* equals the channel's normalized coherent time. During a coherent time (period), i.e., in a frame interval, the channel undergoes quasi-static fading. However, each frame suffers independent fading. $\underline{h}$ is defined by

(5) $$\underline{h} = \left[\underline{h}_1^t \quad \underline{h}_2^t \quad \cdots \quad \underline{h}_{Nr}^t\right]^t$$

Furthermore, $\underline{n}$ is given by

(6) $$\underline{n} = \left[\underline{n}_1^t \quad \underline{n}_2^t \quad \cdots \quad \underline{n}_{N_r}^t\right]^t$$

and A is given by

(7) $$A = \overbrace{\begin{bmatrix} A_N & 0 & \cdots & \cdots & 0 \\ 0 & A_N & 0 & \ddots & \vdots \\ \vdots & 0 & \ddots & 0 & \vdots \\ \vdots & \ddots & 0 & \ddots & 0 \\ 0 & \cdots & \cdots & 0 & A_N \end{bmatrix}}^{N_r}$$

(8) $$A_N = \left\lfloor A_{1,N} \quad A_{2,N} \quad \cdots \quad A_{N_t,N} \right\rfloor$$

(9) $$A_{i,N} = \begin{bmatrix} a_i^{k+1} & a_i^{k} & \cdots & a_i^{k-L+2} \\ a_i^{k+2} & a_i^{k+1} & \cdots & a_i^{k-L+3} \\ \vdots & \vdots & \ddots & \vdots \\ a_i^{k+N} & a_i^{k+N-1} & \cdots & a_i^{k+N-L+1} \end{bmatrix},$$

$$i = 1,2,\cdots,N_t$$

Because keeping the vector form for the channel $\underline{h}$, we design the data matrix as the diagonal matrix. Elements in the diagonal are all identical, it consists of current data and previous data symbols from all transmitter antennas. The length of the later symbols is a function of the channel length *L*.

Eq. (2) indicates that the conditional probability density function (pdf) of the received signal $\underline{r}$, given A and $\underline{h}$, is

(10) $$f(\underline{r}/A,\underline{h}) = \frac{1}{\left(\sigma_n\sqrt{2\pi}\right)^{N_t \cdot N}} \exp\left(\frac{\left|\underline{r} - A\underline{h}\right|^2}{-2\sigma_n^2}\right)$$

where σ_n^2 is the variance of the noise. Let

(11) $$C(A,\underline{h}) = \left|\underline{r} - A\underline{h}\right|^2$$

be the likelihood function. Minimizing this likelihood function with respect to $\underline{h}$ given the data matrix A is utilizing a least squared estimation [8].

(12) $$\frac{\partial C(A,\underline{h})}{\partial \underline{h}} = 2A^H\underline{r} - 2A^H A\underline{h}$$

where superscript H is represented function of transpose and conjugate. If Eq. (12) is set to zero, the ML estimate of $\underline{h}$ will be

(13) $$\underline{\hat{h}}_{ML} = \left(A^H A\right)^{-1} A^H \underline{r}$$

In Eq. (13), we must consider that if $\left(A^H A\right)^{-1}$ does not exist. Actually, $\left(A^H A\right)^{-1} A^H$ is the pseudo-inverse of A and $\left(A^H A\right)$ may be not full rank, the singular value decomposition (SVD) can be used to compute the pseudo-inverse of A.

The new likelihood function by substituting the above $\hat{\underline{h}}_{ML}$ into Eq. (11) is obtained

$$(14) \qquad C(A, \hat{\underline{h}}_{ML}) = \left| \underline{r} - A\left(A^H A\right)^{-1} A^H \underline{r} \right|^2$$

It can be easily proved that minimizing Eq. (14) is equivalent to maximizing Eq. (15) as follows

$$(15) \qquad C'(A) = \underline{r}^H A\left(A^H A\right)^{-1} A^H \underline{r}$$

where calculating the Eq. (15) only need to received signal $\underline{r}$ and transmitted data matrix A without the function of $\underline{h}$.

Our proposed scheme for channel estimation is very clear. From Fig 1. First, we utilize PML algorithm with N pilot symbols to find the $\hat{\underline{h}}'_{ML}$ then put it into the usual VA to acquire the l estimated symbols in sequence estimation. Finally using N pilot symbols and l estimated symbols with PML algorithm again to find the $\hat{\underline{h}}_{ML}$. This estimation will send into the iterative serial concatenated decoder to estimate transmitted data sequence.

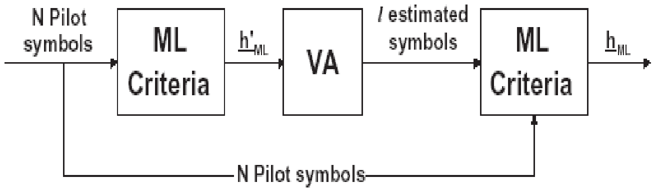

Fig. 1. A block diagram of PML combined with VA for channel estimation with training information.

A.2 *Block least mean square (BLMS) algorithm*

A new way to implement the VA for maximum-likelihood data sequence estimation (MLSE) in a known channel environment and utilize it to derive block adaptive techniques for joint channel and data estimation [4]. The received sequence is fed into the adaptive estimation technique in block of N_b symbols at a time.

In the derivation of the adaptive schemes for the channel state estimation we will assumed that the channel is unknown but static. It match the assumption of ST codes, each transmitted data frame is suffered by quasi-static fading channel. The derivation of this adaptive technique utilizes the ML criterion similar to PML algorithm according to the joint ML estimates of the channel and data are those that maximize the conditional pdf $p(\underline{r} \mid A, \underline{h})$. Therefore, $\hat{\underline{h}}_{ML}$ owned the similar result to the result of PML algorithm that data is assumed to give as follows

$$(16) \qquad \hat{\underline{h}}_{ML} = \left(A^H A\right)^{-1} A^H \underline{r}$$

Based on the above expressions, an iterative procedure was proposed to reach the solution $\hat{\underline{h}}_{ML}$. From Eq. (16), we rewrite it with the function of the number of recursion using the result of this iteration to predict the channel estimation of the next iteration.

$$(17) \qquad \hat{\underline{h}}(l+1) = \left(\hat{A}^H(l)\hat{A}(l)\right)^{-1} \hat{A}^H(l)\underline{r}_l$$

where l is for the lth iteration and $\underline{r}_l$ is for the lth received block within N_b received symbols. Based on the concept of least mean seqare (LMS) [8], we chose to transfer the CSI estimator into one with iterative form as follows,

$$\hat{\underline{h}}(l+1) = \hat{\underline{h}}(l) + « \frac{\partial\left[\ln p\left(r \mid \hat{A}, \underline{h}\right)\right]}{\partial \underline{h}}\Big|_{\underline{h} = \hat{\underline{h}}(l)}$$

$$(18) \qquad = \hat{\underline{h}}(l) + «\hat{A}^H\left[\underline{r} - \hat{A}\hat{\underline{h}}(l)\right]$$

where « is a step-size parameter adjusting the convergence speed. If we can achieve perfect convergence in Eq. (18) then

$$\hat{\underline{h}}(l+1) = \hat{\underline{h}}(l) \Leftrightarrow \hat{A}^H\left[\underline{r} - \hat{A}\hat{\underline{h}}(l)\right] = 0 \Leftrightarrow \hat{\underline{h}}(l)$$

$$(19) \qquad = \left(\hat{A}^H \hat{A}\right)^{-1} \hat{A}^H \underline{r}$$

Therefore, we have the form of adaptive channel estimation technique named block least mean square (BLMS) algorithm.

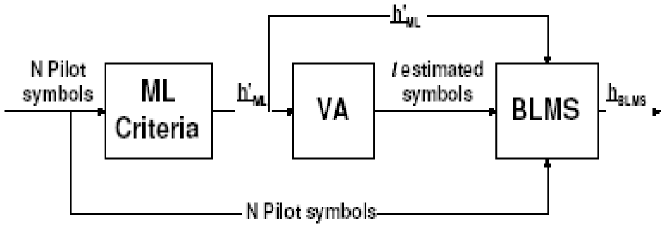

Fig. 2. A block diagram of BLMS combined with VA for channel estimation with training information.

Fig. 2. show a block diagram of BLMS combined with VA for channel estimation. We use ML criterion with N pilot symbols to acquire a tentative channel estimate $\hat{\underline{h}}'_{ML}$ then utilizing channel estimation of ML with VA for l estimated symbols we need. Finally N pilot symbols and l estimated symbols and the tentative channel estimate $\hat{\underline{h}}'_{ML}$ as initial of channel estimation for BLMS algorithm (can speed up the convergence of channel estimation) are all sent into BLMS block for action. According to input information from operation of previous blocks and BLMS algorithm from illustration of this subsection through recursive computations until acceptable convergence of estimation has happened then the final channel estimation $\hat{\underline{h}}_{BLMS}$ is acquired and used for

iterative decoding of serial concatenated ST cods as mentioned in previous chapter for data symbols next.

III. BLIND CHANNEL ESTIMATION

The schemes of channel estimation with training information would waste bandwidth efficiency and power. In this subsection, the blind channel estimation schemes are proposed. We do not use the training information anymore. Our proposed schemes are based on the concept of exhaustive search for finding the reliable information which are served as the approximately training information. Then using the previous proposed schemes to do channel estimation. But the exhausting search needs not only more time but also more computations. In fact, it is not practiced. So our propose is to reduce the complexity of original exhaustive search. The proposed schemes can greatly reduce its complexity and is more efficiency in practice.

A. Branch and Bound Algorithm

The branch and bound algorithm is a general search method. Suppose we wish to minimize a function *f(x)*, where *x* is restricted to some reasonable region (defined by some explicit mathematical constraints). To apply branch and bound, one must have a means of dividing the feasible region of a problem to create smaller subproblems. There must also be a way to compute an upper bound (reasonable solution) for at least some examples; for practical purposes, it should be possible to compute upper bounds for some set of reasonable regions. Next we review the formula about the ML of channel estimation.

The cost function of Eq. (14) can be rewritten as the form of the sum of cost function for each receive antenna as follows,

(20) $$C\left(A,\underline{\hat{h}}_{ML}\right)=\sum_{j=1}^{N_r}\left|\underline{r}-A_N\left(A_N^H A_N\right)^{-1}A_N^H\underline{r}_j\right|^2$$

If the $\left(A_N^H A_N\right)^{-1}$ is not existed. According to the appendix A, the pseudo-inverse of A can be substituted for it and Eq. (20) becomes,

(21) $$C\left(A,\underline{\hat{h}}_{ML}\right)=\sum_{j=1}^{N_r}\left|\underline{r}-A_N A_N^{+}\underline{r}_j\right|^2$$

Before the derivation of our proposed algorithm with branch and bound, we first make some definitions. Let C_{k+d} be the cost function and defined as follows

$$C_{k+d}=\min_{\underline{a}_{k+d+1}^t,\cdots,\underline{a}_{k+N}^t} f\left(\underline{a}_{k+1}^5,\cdots\underline{a}_{k+d}^t,\underline{a}_{k+d+1}^t,\cdots,\underline{a}_{k+N}^t\right)$$

$$\begin{cases}\underline{a}_{k+1}^t,\cdots\underline{a}_{k+d}^t\\ \underline{a}_{k+d+1}^t,\cdots,\underline{a}_{k+N}^t\in C\end{cases}$$

Note that $\underline{a}_{k+1}^t\cdots\underline{a}_{k+d}^t$ are known and fixed. It means that these signals are belonged to the constellation of modulated signals. On the other hand $\underline{a}_{k+d+1}^t\cdots\underline{a}_{k+N}^t$ are distributed in the complex plane. In other words, they could be any complex numbers. According to the combination of the constrained and unconstrained data symbols, the function *f* could be minimized and the value of this function is defined as the cost function. and an important fact is that the cost function is also increased as time increased,

(22) $$C_{k+1}\le C_{k+2}\le C_{k+3}\le\cdots\le C_{k+N}$$

from Eq. (20) with respect to A and h, the cost function is

(23) $$C(A,\underline{h})=\sum_{j=1}^{N_r}\left|\underline{r}_j-A_{\underline{N}}\underline{h}_j\right|^2$$

$\underline{r}_j$, A_N and $\underline{h}_j$ are the same as before. From the above definition, we can separate the cost function into two parts, the known and fixed part and the unconstrained part.

$$\left|\underline{r}-A_N\underline{h}_j\right|^2=\left\|\begin{bmatrix}\underline{r}_j^p-A_N^p\underline{h}_j\\ \underline{r}_j^q-A_N^q\underline{h}_j\end{bmatrix}\right\|^2$$
$$=\left|\underline{r}_j^p-A_N^p\underline{h}_j\right|^2+\left|\underline{r}_j^q-A_N^q\underline{h}_j\right|^2$$

where the superscript p is for the known and fixed part from time $k+1$ to $k+d$. where superscript q is for the unconstrained part from time $k+d+1$ to $k+N$.

Because the cost function of the known and fixed part are exactly known, we will focus on the minimization of the cost function of the unconstrained part.

(24) $$\frac{\partial}{\partial\left(A_N^q\underline{h}_j\right)}\left|\underline{r}_j^q-A_N^q\underline{h}_j\right|^2=0\Rightarrow A_N^q\underline{h}_j=\underline{r}_j^q$$

Eq. (24) shows the ML method of the cost function of the unconstrained part. By setting the partial derivative of $\left|\underline{r}_j^q-A_N^q\underline{h}_j\right|^2$ with respect to $A_N^q\underline{h}_j$ to zero, we can acquire the ML solution for this cost function.

The result of this solution looks like very straightforward, but we have made an very important assumption that the data symbols of the unconstrained part can be any values in the complex plane. Therefore, theoretically we can find a data sequences such that the cost function of the unconstrained part is zero. Finally, the cost function of Eq. (23) can be determined only by the cost function of the known and fixed part is shown,

(25) $$C(A,\underline{h})=\sum_{j=1}^{N_r}\left|\underline{r}_j-A_N\underline{h}_j\right|^2=\sum_{j=1}^{N_r}\left|\underline{r}_j^p-A_N^p\underline{h}_j\right|^2$$

Fig. 3 shows the tree of all possible sequences which can be used to explain the branch and bound algorithm. First, we randomly choose a data sequences of length N then find a reference cost function of it. Second from the initial node, data sequences prior to time $k+2$

are set to be known and we can determine the cost function by some candidate data sequences at time $k+1$ is C_{k+1}. If C_{k+1} is greater than the reference cost function, according to Eq. (22) the corresponding cost function with time will always be greater than the reference cost function. Therefore, we can immediately determine that the further computation is not necessary beyond this node. This step will reduce the number of computation greatly and make this blind algorithm more practicable.

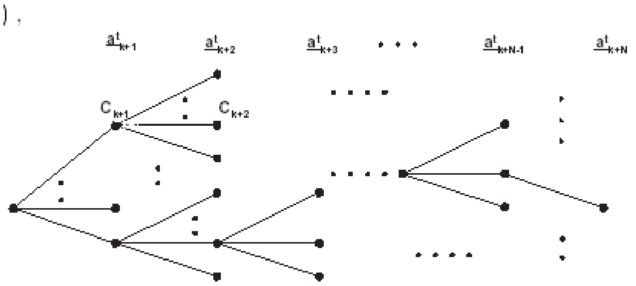

Fig. 3. A tree of all possible sequences.

But if C_{k+1} is less than the reference cost function, the data at time $k+2$ will be added and now data sequences prior to time $k+3$ is set to be known. The cost function C_{k+2} is presented and compared with the reference cost function. Repeating the same procedure until at time $k+N$ then compare the C_{k+N} with the reference function. If it is less than the reference cost function, the reference cost function is replaced by C_{k+N} or else the cost function is kept and we return back to the initial node to try another data sequences for the same procedure until all possible candidate data sequences are searched.

After all possible candidate data sequence are searched, the final data sequence is served as our simulated pilot symbols. We can utilize the proposed scheme suitable for the system with training information again to acquire the channel estimation, then iterative decoding procedures are performed as mentioned before. Similarly, Fig. 4 and Fig. 5. show the blind channel estimation diagram combined with the ML and BLMS algorithm, individually.

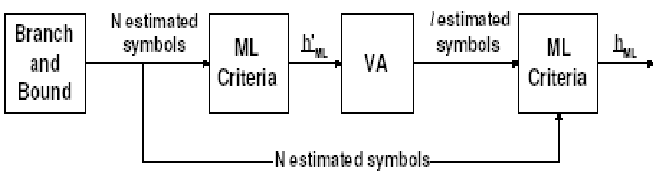

Fig. 4. A block diagram of PML combined with VA for channel estimation with branch and bound algorithm.

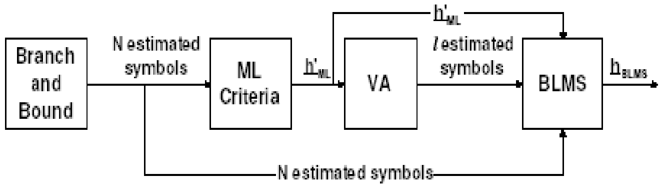

Fig. 5. A block diagram of BLMS combined with VA for channel estimation with branch and bound algorithm.

IV. PERFORMANCE OF SERIAL CONCATENATED SPACE-TIME BLOCK CODES

In the simulation of the serial concatenated ST block codes, we will show the performance result of PML and BLMS algorithm with training information and with branch and bound algorithm (blind channel estimation). The interleaver size and frame length is 130. The outer code, convolutional code, has the (2,1,3) form. The length of multipath fading channel is 2 and quasi-static fading is assumed during each frame symbol. Scheme with training information, 10 pilot symbols are sent and VA is utilized to acquire 20 estimated data symbols. Then all 30 data symbols are performed by PML or BLMS. In branch and bound algorithm, the reference cost function chooses the reliable one from four reference values. 10 pilot symbols is replaced by utilizing branch and bound algorithm to acquire then the following procedure is the same as scheme with training information.

Fig. 6 and Fig. 7 show the FER and BER of serial concatenated ST block code (Tx:2, Rx:1) with perfect CSI and BLMS scheme about VA over slow and frequency selective fading channel. For the ST block codes its decoding will eliminate the multipath channel effect, therefore, the relationship between symbol by symbol will loosen. The effect of extrinsic information of the iterative decoding has smaller influence on it so the performance the serial concatenated ST block codes has unapparent improvement by iterative decoding procedure.

The performance of BLMS scheme is better than it of PML scheme. Because the number of training information symbols is limited for bandwidth efficiency, the adaptive scheme based on BLMS will track the change of channel and noise and converge to find the better solution than find it only with ML algorithm.

Fig. 10 shows the number of computation with branch and bound algorithm for ST block codes (Tx:2,Rx:1) over slow and frequency selective fading channel. Giving only one reference value or choosing reliable one from four reference values, the latter has fewer number of computations but basically the performances of both are the same.

There are many related issues remain to be solved. For examples, the optimal detector for ST block coded system in frequency selective fading is not available, more efficient blind detectors are needed. Moreover, we have been dealing with static channels only. The effect of channel dynamic and the capability of each proposed scheme to track the channel variation deserve further investigation.

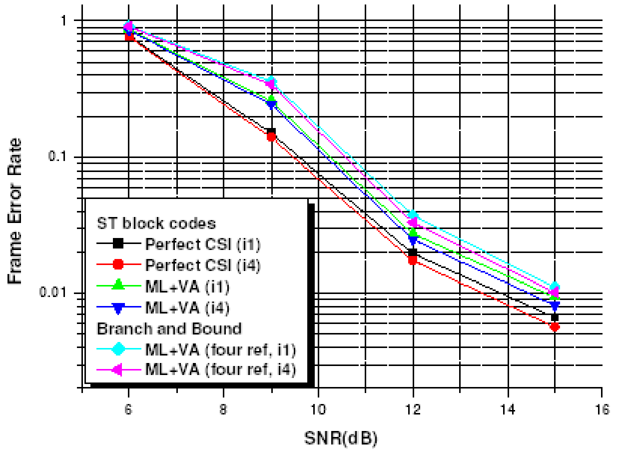

Fig. 6. FER of serial concatenated ST block code (Tx:2, Rx:1) with perfect CSI and PML scheme about VA over slow and frequency selective fading channel.

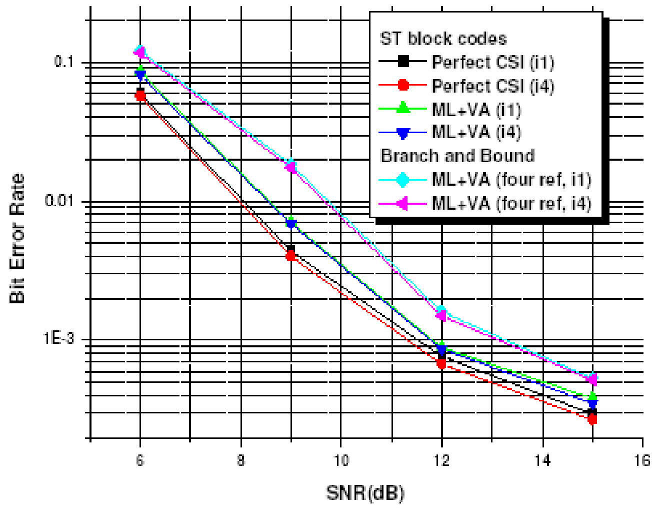

Fig. 7. BER of serial concatenated ST block code (Tx:2,Rx:1) with perfect CSI and PML scheme about VA over slow and frequency selective fading channel.

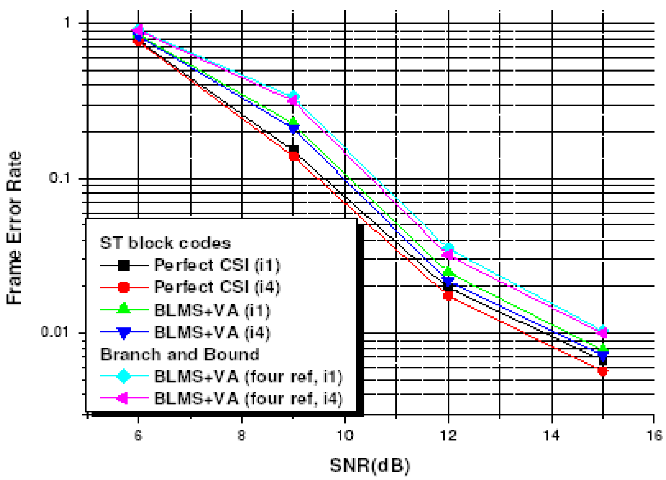

Fig. 8. FER of serial concatenated ST block code (Tx:2, Rx:1) with perfect CSI and BLMS scheme about VA over slow and frequency selective fading channel.

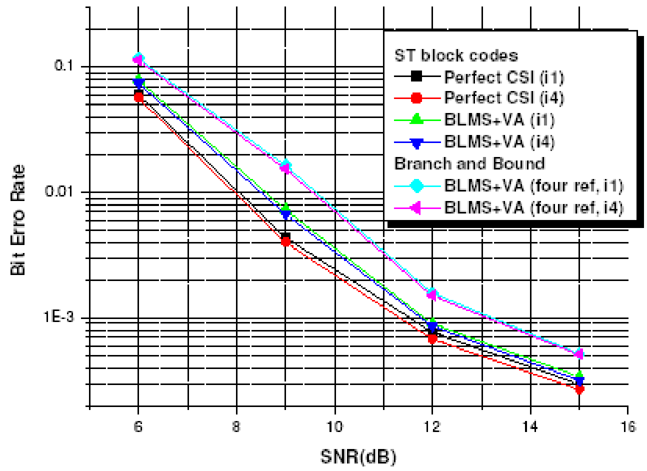

Fig. 9. BER of serial concatenated ST block code (Tx:2,Rx:1) with perfect CSI and BLMS scheme about VA over slow and frequency selective fading channel.

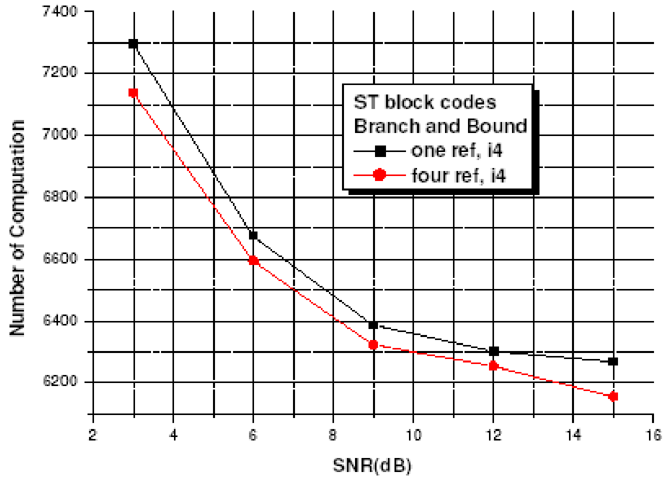

Fig. 10. Number of computation with branch and bound algorithm for ST block codes (Tx:2,Rx:1) over slow and frequency selective fading channel.

V. CONCLUSIONS

In this paper, we investigate the receiver structures for serially-concatenated codes with a ST block code as the inner codes and a convolutional code as the outer code. The decoder is based on ML and SOVA algorithms for iterative decoding of the ST block and convolutional codes. We proposed several channel estimators for this systems. They are divided into two categories, namely, the pilot-assisted estimators and the blind estimators. The former category includes two algorithms: PML and BLMS methods. Both methods are used in conjunction with VA to reduce the number of required pilots. The blind estimators have structures similar to those of non-blind estimators. Basically they add an additional blind data detection step and use the estimated short sequence to act as training symbols. Using the proposed branch-and-bound method, our blind estimators do achieve satisfactory performance with moderate complexity.

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