A Low Complexity Blind Data Detector for OFDM Systems

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Abstract—**A low-complexity blind data detector is proposed in this paper for orthogonal frequency division multiplexing (OFDM) systems, where the generalized likelihood ratio test (GLRT) approach is adopted. Traditional GLRT data detector (GDD) can be viewed as a combinatorial optimization problem, which suffers from prohibitively high computational complexity. The proposed scheme reduces the search space by exploiting the phase ambiguity. In addition, the computational complexity is further reduced by properly decoupling the GDD into several sub-group GDDs (SGDD). Mathematical analysis is adopted to evaluate the complexity of the proposed scheme. Simulation experiments are conducted to verify the system performance.**

Keywords-Orthogonal frequency division multiplexing (OFDM), blind data detection, generalized likelihood ratio test (GLRT), phase ambiguity.

I. Introduction

Orthogonal frequency division multiplexing (OFDM) has emerged as an effective multicarrier transmission scheme for wireless communication over frequency selective fading channels [1-2]. OFDM technique converts the frequency selective fading channel into a series of flat fading channels by transmitting information symbols in parallel.

In OFDM systems, the channel estimation is generally performed with the aid of pilot symbols, which substantially reduce the bandwidth utilization. Therefore the blind channel estimation methods receive considerable attention [3-5]. Unfortunately, blind channel estimation is usually restricted by the channel coherent time since it requires a large number of symbols to ensure the accuracy of statistical properties. Hence, the joint channel estimation and data detection, which requires fewer pilots and can be accomplished in an OFDM symbol, is of substantial interest [6-11]. In [8-9, 11], this problem was transformed into a data detector via the generalized likelihood ratio test (GLRT) approach [12]. It is worth noting that the GLRT data detector (GDD) proposed in [9] is a combinatorial optimization problem, which in general can be solved by exhaustively searching the finite space. However, the exhaustive search is impractical since the search space is exponentially proportional to the number of subcarriers, which is generally greater than 100 in modern communication systems. In addition, although the sphere decoding (SD) algorithm can be exploited to reduce the computational complexity, it still requires prohibitively computational complexity [10]. Therefore, a novel methodology is proposed in this paper to reduce the search space of GDD.

This paper proposes a low-complexity blind data detector for OFDM systems, where the search space of GDD is substantially reduced. In the proposed scheme, the GDD is decomposed into several subgroup-GDDs (SGDD), a similar approach to that proposed by [11]. It is proved in this study that a unique solution of SGDD is guaranteed only if the number of sub-carrier in an SGDD is greater than the channel order. However, as opposed to the SGDD proposed in [11], where a pilot sub-carrier is appended to each subgroup to resolve the phase ambiguity, the proposed scheme taking advantage of phase ambiguity to decrease the computational complexity. It is demonstrated that, by using the phase ambiguity, the search space of GDD can be reduced to $1/M$ for the case of M-PSK modulation. It is also proved that the sub-carriers of a given subgroup should be equally spaced to obtain a low computational complexity.

The rest of this paper is organized as follows. Section II describes the system architecture and the signal model. Section III introduces the GDD method. Section IV develops the structure of SGDD, where the phase ambiguity is exploited to reduce the search space. Section V presents the simulation results. Finally, some concluding remarks are provided in Section VI.

II. System Architecture And Signal Model

In this paper, we consider an OFDM system with N sub-carriers. In addition, the data are M-PSK modulated, where the input data stream is mapped onto a finite constellation $\mathbf{\Phi}$ with equal amplitude. Note that the assumption of M-PSK modulation is made for introductory purpose. The proposed scheme can be adopted by various modulation schemes.

Given the $N\times 1$ column vector $\mathbf{S}$, which denotes the frequency domain OFDM signal, the corresponding time domain OFDM signal $\mathbf{s}$, excluding the cyclic prefix (CP), is obtained by applying an inverse discrete Fourier transform (IDFT) operation, i.e.

$$\mathbf{s} = \frac{1}{N}\mathbf{F}^{H}\mathbf{S}, \tag{1}$$

where $\mathbf{F}$ is the $N\times N$ discrete Fourier transform (DFT) and the $(m,\ n)$th entry has the form $F(m,n)=\exp(-j2\pi mn/N)$. In addition, $(\bullet)^{H}$ denotes the Hermitian operation. The cyclic prefix (CP), a copy of the last N_{CP} samples, is appended at the head of each OFDM symbol.

An assumption is made in this work that the channel is quasi-stationary, i.e. the channel remains static over an entire OFDM symbol. The discrete-time channel impulse response (CIR) is given by

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$$\mathbf{h}=[h(0),h(1),...,h(L-1)]^T, \tag{2}$$

where L is the number of resolvable paths. The channel taps $h(l)$, $0 \le l \le L-1$, are assumed to be independent complex Gaussian distributed random variables with zero mean and a variance $\sigma^2_{h(l)}$. The channel frequency response $\mathbf{H}$ has the following form,

$$\mathbf{H}=\left[H(0),H(1),\ldots,H(N-1)\right]^T \equiv \mathbf{Gh}, \tag{3}$$

where $\mathbf{G}$ is an $N\times L$ matrix constructed by the first L columns of $\mathbf{F}$. Furthermore, it is assumed that the length of CP is larger than the channel order.

At the receiver, it is assumed that the perfect synchronization is achieved. Therefore, the received signal is free from inter-carrier interference (ICI) and inter-symbol interference (ISI). After the CP removal and DFT operation, the received OFDM symbol in frequency domain is given by:

$$\mathbf{Y}=\mathbf{S}_D\mathbf{H}+\mathbf{W}=\mathbf{S}_D\mathbf{Gh}+\mathbf{W}, \tag{4}$$

where $\mathbf{S}_D$ represents a diagonal matrix with vector $\mathbf{S}$ on the diagonal, and each element in the $N\times 1$ column vector $\mathbf{W}$ represents the complex white Gaussian noise with zero mean and a variance σ^2_W.

III. GLRT Data Detector

In this section, the architecture of GLRT data detector (GDD) is briefly described. According to (4), the maximum likelihood (ML) estimator is given by:

$$\left(\hat{\mathbf{S}},\hat{\mathbf{h}}\right)=\underset{\tilde{\mathbf{S}}\in\mathbf{\Phi}^N,\tilde{\mathbf{h}}\in C^L}{\arg\min}\left\|\mathbf{Y}-\tilde{\mathbf{S}}_D\mathbf{G}\tilde{\mathbf{h}}\right\|^2, \tag{5}$$

where $\tilde{\mathbf{S}}\in\mathbf{\Phi}^N$ denotes an $N\times 1$ vector with elements obtained from the finite constellation $\mathbf{\Phi}$, and $\tilde{\mathbf{h}}\in C^L$ denotes an $L\times 1$ complex vector. If the data sequence $\mathbf{S}$ and the channel order L are given, the least square (LS) solution of CIR can be shown as

$$\hat{\mathbf{h}}=\left[(\mathbf{S}_D\mathbf{G})^H(\mathbf{S}_D\mathbf{G})\right]^{-1}(\mathbf{S}_D\mathbf{G})^H\mathbf{Y}. \tag{6}$$

From (5) and (6), the data detector is given by:

$$\begin{aligned}\hat{\mathbf{S}}&=\underset{\tilde{\mathbf{S}}\in\mathbf{\Phi}^N}{\arg\min}\,\mathbf{Y}^H\tilde{\mathbf{S}}_D\left[\mathbf{I}_N-\mathbf{G}\left(\mathbf{G}^H\mathbf{G}\right)^{-1}\mathbf{G}^H\right]\tilde{\mathbf{S}}_D^H\mathbf{Y}\\&=\underset{\tilde{\mathbf{S}}\in\mathbf{\Phi}^N}{\arg\min}\,\tilde{\mathbf{S}}^T\mathbf{Y}_D^H\left[\mathbf{I}_N-\mathbf{G}\left(\mathbf{G}^H\mathbf{G}\right)^{-1}\mathbf{G}^H\right]\mathbf{Y}_D\tilde{\mathbf{S}}^*,\end{aligned} \tag{7}$$

where $\mathbf{I}_N$ is an $N\times N$ identity matrix. The detector in (7) is known as the GDD [9].

We emphasize that the GDD, as well as other blind detectors, suffers from the problem of phase ambiguity, which is described in *Theorem 1*.

Theorem 1: The optimal solution for GDD is not unique since both $\mathbf{S}_r$ and $e^{j\phi}\mathbf{S}_r$ achieve the same objective value, where ϕ is arbitrary real value.

The proof of *Theorem 1* can be found in [9, 11]. In addition, the phase ambiguity problem is generally solved by adding a pilot tone or by using the superimposed training scheme [9].

IV. GLRT Data Detector with Reduced Search Space

From (7), the dimension of the search space is *N* since all the subcarriers are jointly detected. Hence, in the case of *M*-PSK modulation, GDD can be regarded as an *M*-way partitioning problem on a set of *N* elements [13], which can be solved by exhaustively searching over all M^N possible data sequences. Although an exhaustive search of the solution space yields an optimal solution, the solution space grows exponentially with the modulation order *M* and the number of sub-carriers *N*. Therefore, GDD is in general too complicated for practical implementation. For this reason, it is attractive to reduce the search space of GDD.

In this section, we firstly demonstrate two properties of GDD, showing that the search space of GDD is reducible. Subsequently, these two properties are adopted to develop a low-complexity searching algorithm. Finally, the computational complexity of the proposed scheme is analyzed.

A. Exploiting phase ambiguity to reduce search space

As mentioned above, the phase ambiguity of GDD is generally regarded as a drawback since it requires additional pilot symbols and the bandwidth efficiency is thus reduced. However, it will be demonstrated subsequently that the phase ambiguity can be utilized to reduce the search space of GDD.

Theorem 1 reveals that both $\mathbf{S}$ and $e^{j\phi}\mathbf{S}$ achieve the same objective value, where ϕ is an arbitrary real value. Given an arbitrary data sequence $\mathbf{S}'$, let's define a family of sequences $T(\mathbf{S}')$:

$$T(\mathbf{S}')\triangleq\left\{\mathbf{U}\mid\mathbf{U}=e^{j\phi_m}\mathbf{S}',\phi_m=2\pi m/M,m=0,1,\cdots M-1\right\} \tag{8}$$

We note that the size of family depends on the modulation order *M*. By defining the family of sequences, the whole search space is partitioned into $M^N/M=M^{N-1}$ disjoint families. In addition, all the sequences in $T(\mathbf{S}')$ have the same objective value because of the phase ambiguity described above. In this paper, $\mathbf{S}'$ is designated as the representative sequence of the family $T(\mathbf{S}')$. In addition, the collection of all the representative sequences is denoted as $\theta\left(\mathbf{\Phi}^N\right)$. Therefore, instead of assessing the entire space $\mathbf{\Phi}^N$, which consists of M^N possible solutions, the detector only needs to evaluate the subset $\theta\left(\mathbf{\Phi}^N\right)$, which consists of M^{N-1} representative sequences. Consequently, by exploiting the phase ambiguity, the size of search space is reduced to $1/M$ of the original search space without any performance loss. It is noted that a reference pilot is still required in order to make the final decision within the family $T(\mathbf{S}')$.

B. Subgroup-GDD with reduced search space

As above mentioned, the search space is exponentially proportional to the number of sub-carriers. This motivates us to break the GDD into several subgroup-GDDs (SGDDs). In this paper, the N subcarriers are partitioned into R subgroups, each having $N_{SG} \equiv N/R$ subcarriers. The sub-carrier index set of the rth subgroup is denoted as I_r, $r = 0,1,...,R-1$. From (4), the received signal for the rth subgroup is given by:

$$\mathbf{Y}_r = \mathbf{S}_{r,D}\mathbf{H}_r + \mathbf{W}_r = \mathbf{S}_{r,D}\mathbf{G}_r\mathbf{h} + \mathbf{W}_r, \tag{9}$$

where $(\bullet)_r$ is a sub-matrix/vector which contains the corresponding rows of the original matrix/vector according to I_r. Therefore, the objective function given in (7) becomes a suboptimum data detector, as denoted by:

$$\begin{aligned}\hat{\mathbf{S}}_r &= \underset{\tilde{\mathbf{S}}_r \in \mathbf{\Phi}^{N_{SG}}}{\arg} \min \left\| \mathbf{Y}_r - \tilde{\mathbf{S}}_{r,D}\mathbf{G}_r\left[\mathbf{G}_r^H\mathbf{G}_r\right]^{-1}\mathbf{G}_r^H\tilde{\mathbf{S}}_{r,D}^H\mathbf{Y}_r \right\|^2 \\ &= \underset{\tilde{\mathbf{S}}_r \in \mathbf{\Phi}^{N_{SG}}}{\arg} \min \tilde{\mathbf{S}}_r^T\mathbf{Y}_{r,D}^H\left[\mathbf{I}_{N_{SG}} - \mathbf{G}_r\left(\mathbf{G}_r^H\mathbf{G}_r\right)^{-1}\mathbf{G}_r^H\right]\mathbf{Y}_{r,D}\tilde{\mathbf{S}}_r^*.\end{aligned} \tag{10}$$

Note that each subgroup associates an SGDD. In specific, the GDD is decoupled into R SGDDs. The overall search space of the R SGDDs is given by $R \cdot M^{N/R}$. It is worthy of note that the overall search space of the R SGDDs exponentially decreases with $R = N/N_{SG}$. Therefore, a small N_{SG} is required to obtain a smaller search space. However, it is demonstrated in *Theorem 2* that N_{SG} is lower bounded by the channel order L to guarantee a unique solution of each SGDD.

Theorem 2: The unique solution of each SGDD is guaranteed if the dimension N_{SG} of the suboptimum problem is lower bounded by the order of CIR, i.e. $N_{SG} > L$.

Proof: In order to guarantee a unique solution of each SGDD, it is required that the $L \times L$ matrix $\left(\mathbf{G}_r^H\mathbf{G}_r\right)^{-1}$ exists. In other words, $\mathbf{G}_r^H\mathbf{G}_r$ needs to be full rank. Because

$$\begin{aligned} rank\left(\mathbf{G}_r^H\mathbf{G}_r\right) &\le \min\left(rank\left(\mathbf{G}_r\right), rank\left(\mathbf{G}_r^H\right)\right) \\ &= rank\left(\mathbf{G}_r\right) \le \min\left(N_{SG}, L\right),\end{aligned} \tag{11}$$

$\mathbf{G}_r$ is required to be full column rank, i.e. $N_{SG} \ge L$.

When $N_{SG} = L$, $\mathbf{G}_r$ is a square matrix with full rank. In this case, $\left[\mathbf{I}_{N_{SG}} - \mathbf{G}_r\left(\mathbf{G}_r^H\mathbf{G}_r\right)^{-1}\mathbf{G}_r^H\right] = \mathbf{0}$. As a consequence, the detection results of (10) are always zero, a trivial result, for every data sequence when $N_{SG} = L$. We therefore conclude that the dimension N_{SG} of SGDD has to be larger than the order of CIR, i.e. $N_{SG} > L$, to obtain a non-trivial result. □

It can be observed from (10) that the complexity of inverse matrix operation taken in SGDD is significantly decreased if $\mathbf{G}_r^H\mathbf{G}_r$ is a diagonal matrix. It is proved in *Theorem 3* that the $\mathbf{G}_r^H\mathbf{G}_r$ is a diagonal matrix if and only if the sub-carriers in each subgroup are equally-spaced.

Theorem 3: $\mathbf{G}_r^H\mathbf{G}_r$ is a diagonal matrix if and only if the sub-carriers in each subgroup are equally-spaced.

Proof:

$(\rightarrow)$ Let $\boldsymbol{\psi}_r \equiv \mathbf{G}_r^H\mathbf{G}_r$. If $\boldsymbol{\psi}_r$ is a diagonal matrix, i.e. the (m, n)th entry of $\boldsymbol{\psi}_r$, $\psi_r(m,n) = \sum_{v=1}^{N_{SG}} e^{j\frac{2\pi(I_r(v))(m-n)}{N}} = 0$ for $m \ne n$, then the N_{SG} elements in summation have to be equally-spaced on the unit circle. Therefore, the sub-carrier index set I_r of each subgroup are equally-spaced.

$(\leftarrow)$ The proof is straightforward and then omitted due to limited space. □

We therefore consider in this investigation the equally-spaced subcarrier partition, as shown in the following equation:

$$I_r \triangleq \left\{u \mid u = qR + r, 0 \le q \le N_{SG} - 1\right\}, \quad 0 \le r \le R-1. \tag{12}$$

C. Proposed two-stage SGDD

From the above discussions, the computational complexity of GDD can be reduced by utilizing the phase ambiguity property and by decoupling GDD into R SGDDs, where the number of sub-carriers in each subgroup should be greater than the channel order. Therefore, in our proposed scheme, the N subcarriers are partitioned into R disjoint subgroups according to the following equation,

$$I_r^{\text{Prop.}} = \begin{cases}\left\{u \mid u = qR + r, q = 0,1,...,N_{SG} - 1\right\} & r = 0,1,...,R-2, \\ \left\{u \mid u = qR + r, q = 0,1,...,N_{SG} - 2\right\} & r = R-1.\end{cases} \tag{13}$$

In addition, a single pilot P is allocated on the last subcarrier.

The proposed data detection process is comprised of two stages. In the first stage, the family of data sequences that achieves the minimum objective value is identified. We recall that all the data sequences of a given family have the same objective value because of the phase ambiguity. In the second stage, the phase ambiguity is resolved with the aid of pilot symbols. To this purpose, the pilot symbol located at the last sub-carrier, i.e. the (N-1)th sub-carrier, is appended to all the sub-groups for data detection. Therefore, the sub-carrier indexes of the rth sub-group are now denoted as

$$I_{rp}^{\text{Prop.}} \triangleq I_r^{\text{Prop.}} \cup \left\{N-1\right\}, \quad r = 0,1,...,R-1. \tag{14}$$

In summary, the data detection processes can be formulated as follows:

Stage 1:

$$\breve{\mathbf{S}}_r = \underset{\tilde{\mathbf{S}}_r \in \theta\left(\mathbf{\Phi}^{\left|I_r^{\text{Prop.}}\right|}\right)}{\arg} \min \tilde{\mathbf{S}}_r^T\mathbf{Y}_{r,D}^H\left[\mathbf{I}_{\left|I_r^{\text{Prop.}}\right|} - \mathbf{G}_r\left(\mathbf{G}_r^H\mathbf{G}_r\right)^{-1}\mathbf{G}_r^H\right]\mathbf{Y}_{r,D}\tilde{\mathbf{S}}_r^*, \tag{15}$$

$$T\left(\breve{\mathbf{S}}_r\right) = \left\{\mathbf{U} \mid \mathbf{U} = e^{j\phi_m}\breve{\mathbf{S}}_r, \phi_m = 2\pi m/M, m = 0,1,\cdots M-1\right\}. \tag{16}$$

Stage 2:

$$\hat{\mathbf{S}}_r = \underset{\tilde{\mathbf{S}}_r \in T\left(\breve{\mathbf{S}}_r\right)}{\arg} \min \left[\tilde{\mathbf{S}}_r^T \; P\right]\mathbf{Y}_{rp,D}^H\left[\mathbf{I}_{\left|I_{rp}^{\text{Prop.}}\right|} - \mathbf{G}_{rp}\left(\mathbf{G}_{rp}^H\mathbf{G}_{rp}\right)^{-1}\mathbf{G}_{rp}^H\right]\mathbf{Y}_{rp,D}\left[\tilde{\mathbf{S}}_r^T \; P\right]^H \tag{17}$$

It is noted that each SGDD in (15) only evaluates the objective value of the representative data sequence in each family and then the family with the minimum objective value is selected. The family of data sequences identified in the first stage is denoted as $T\left(\breve{\mathbf{S}}_r\right)$, $r=0,1,\ldots,R-1$. In the second stage of data detection given in (17), the decision $\hat{\mathbf{S}}_r$ is made by choosing a data sequence from the family of data sequences $T\left(\breve{\mathbf{S}}_r\right)$ with the aid of pilot symbol. The final decision is a data sequence $\hat{\mathbf{S}}$ of length N, which is basically the concatenation of the $\hat{\mathbf{S}}_r$, r=0,1,…, R-1, obtained in the second stage.

It is noted that the sub-carrier partition is also adopted in [11], where the partition rule is given by:

$$I_r^{[11]}=\begin{cases}\{u \mid u=qR+R-1, 0\le q\le N_{SG}-1, r=R-1\},\\ \{u \mid u=qR+r, 0\le q\le N_{SG}-1, 0\le r\le R-2\}\cup\{N-1\},\end{cases} \tag{18}$$

where the (N-1)th subcarrier is the pilot subcarrier. It is seen that the pilot sub-carrier is included in all the subgroups to resolve the phase ambiguity. Apparently, the data detector proposed in [11] has a higher computational complexity as will be analyzed in the following sub-section.

D. Analysis of computational complexity

The computational complexity of the proposed data detector and that proposed in [11] are analyzed in the following:

1. The data detector proposed in [11]:

According to the subcarrier partition shown in (18), the total computational complexity is the summation of the two following parts:

i) For $r=R-1$: the total number of possible data sequences is $M^{N_{SG}-1}$;

ii) For $r=0,1,\cdots,R-2$: the size of search space is $M^{N_{SG}}$.

Hence the computational complexity of the data detector proposed in [11] is:

$$O\left(M^{N_{SG}-1}+(R-1)M^{N_{SG}}\right). \tag{19}$$

2. The proposed two-stage data detector:

i) The first stage:

a) For $r=R-1$: the search space is given by $M^{N_{SG}-1}/M=M^{N_{SG}-2}$ because $\left|I_{R-1}\right|=N_{SG}-1$;

b) For $r=0,1,\cdots,R-2$: The search space is given by $M^{N_{SG}}/M=M^{N_{SG}-1}$;

ii) The second stage: there are M possible solutions in each of the R SGDDs.

Therefore, the total complexity is given by:

$$O\left(M^{N_{SG}-2}+(R-1)M^{N_{SG}-1}+RM\right). \tag{20}$$

A comparison between (19) and (20) indicates that the proposed data detector has a lower computational complexity since the phase ambiguity is exploited to reduce the search space.

In addition, it is well known that the sphere decoding (SD) algorithm is more efficient than exhaustive search and can be adopted by the searching problem presented in (7). It should be emphasized that the proposed scheme can also be combined with the SD algorithm, making the proposed scheme more attractive in practical implementation.

V. SIMULATION RESULTS

The performance of the proposed two-stage data detection process is evaluated by conducting simulation experiments. The data of the investigated system are QPSK modulated and the pilot $P=(1+j)$ is adopted. In addition, the multi-path channel has a length of L=4 and the variances of various channel taps are given by $\sigma^2_{h(l)}=\gamma_l\exp(-\alpha l)$, $l=0,\ldots,L-1$, where $\alpha=0.1$ and γ_l is the normalization coefficient such that $\sum_{l=0}^{L-1}\sigma^2_{h(l)}=1$. The receiver is assumed to have perfect knowledge of the channel length.

The first set of simulation experiments was conducted to evaluate the bit error rate (BER) performance of the proposed two-stage data detector and the traditional GDD [9]. Since the complexity of traditional GDD is exponentially proportional to the number of sub-carriers, it is assume that the investigated OFDM system has N=16 sub-carriers and each OFDM symbols has a CP of length 4. As for the proposed scheme, the sub-carriers are partitioned into 2 subgroups and the size of each subgroup is N_{SG}=8. In addition, the pilot is transmitted on the 15th sub-carrier. The results presented in Fig. 1 demonstrate that the traditional GDD scheme has a better BER performance since it provides the global optimum solution at the sacrifice of computational complexity. The case of perfect channel state information (CSI) shown in Fig.1 can be treated as the BER lower bound. Note that a very low complexity technique named cyclic ML detection can be adopted after GLRT data detection for performance enhancement. However, this is beyond our scope of this paper. The reader interested may refer to [11] and the references therein.

In the second set of simulation experiments, the BER performance of the proposed two-stage data detector is compared with that of [11]. In this simulation, the OFDM system is assumed to have N=64 sub-carriers and the pilot is allocated on the 63th sub-carrier. In addition, the sub-carriers are partitioned into 4 or 8 subgroups, with N_{SG}=16 or 8, respectively, for both schemes. The simulation results are delineated in Fig. 2. It can be seen that the performance of the proposed two-stage data detector is basically the same as that proposed in [11], or the difference is negligible. In addition, Fig. 2 reveals that the performance of both schemes improves with the increase of N_{SG}, the number of sub-carriers in each subgroup. However, an increase of N_{SG} also increases the computational complexity.

The size of search space as a function of the number of sub-carriers, N, is depicted in Fig. 3, where N_{SG} =16. It can be seen in Fig. 3(a) that the size of search space of traditional GDD exponentially increases with N. On the other hand, search space of the proposed scheme and [11] only linearly increases with N

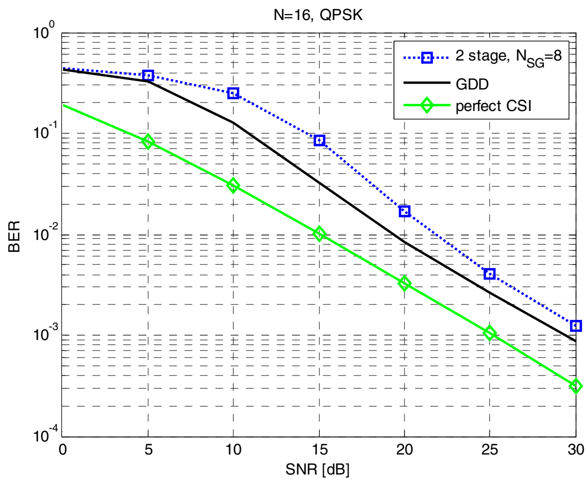

Figure 1: BER comparison between GDD [9] and the proposed two-stage data detector.

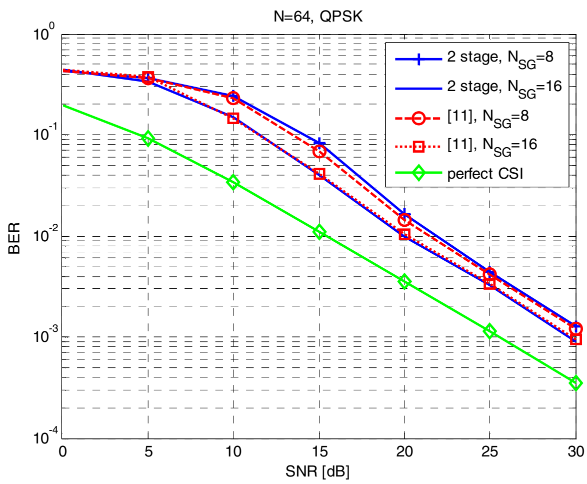

Figure 2: BER comparison of [11] and the proposed two-stage data detector under different size of subgroup.

because of the sub-carrier partition. In addition, Fig. 3(b) demonstrates that the proposed method enjoys a much smaller search space with the help of phase ambiguity.

VI. Conclusition

A two-stage sub-optimal GLRT data detector is investigated in this paper. The proposed scheme takes advantages of the phase ambiguity and sub-carrier partition to decrease the search space. It is shown that the sub-carriers within a subgroup should be equally spaced to obtain a low computational complexity. It is also shown that the number of sub-carriers in each subgroup should be greater than the channel order to guarantee a unique solution. Computer simulations verify that the proposed data detector has the same BER performance as that of [11]. However, the proposed scheme has a much lower computational complexity.

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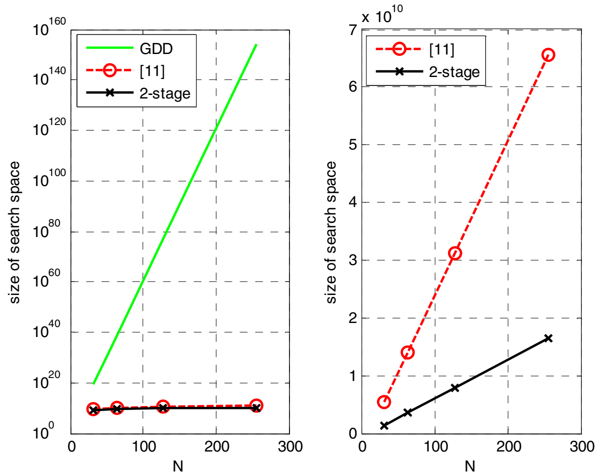

Figure 3: The size of searching space versus the total number of subcarrier *N*. (a): Comparison of GDD, [11], and the proposed scheme in log scale. (b): Comparison of [11] and the proposed scheme in normal scale.

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