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Digital Joint Equivalent Channel Hybrid Precoding for MillimeterWave Massive MIMO Systems

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Abstract

Aiming at the problem that the spectral efficiency of hybrid precoding (HP) is too low in the current millimeter-wave (mmWave) massive multiple input multiple output (MIMO) system, this paper proposes a digital joint equivalent channel hybrid precoding algorithm. Based on the introduction of digital encoding matrix iteration. First, the objective function is expanded to obtain the relation equation, and the pseudo-inverse iterative function of the analog encoder is derived by using the pseudo-inverse method, which solves the problem of greatly increasing the amount of computation caused by the lack of rank of the digital encoding matrix, and reduces the overall complexity of hybrid precoding. Secondly, the analog coding matrix and the millimeter-wave sparse channel matrix are combined into an equivalent channel, and then the equivalent channel is subjected to Singular Value Decomposition (SVD) to obtain a digital coding matrix, and then the derived pseudo-inverse iterative function is used to iteratively regenerates the simulated encoding matrix. The simulation results show that the proposed algorithm improves the system spectral efficiency by 10∼20%compared with other algorithms, and the stability is also improved.

Keywords: MmWave,massive MIMO, hybrid precoding, singular value decompositing, equivalent channel.

1 Introduction

With the rapid development of information industry technology represented by 5G, various electronic smart devices have gradually entered people's lives. However, with the increase of device interfaces, the original bloated frequency band has been unable to bear the increasing data transmission demand, so that the millimeter wave (mmWave) with higher frequency band has gradually entered the field of vision [1-3].

However, the millimeter wavelength of 1-10mm often leads to high loss, and massive MIMO is introduced to deal with this situation. Massive MIMO utilizes the multiple gains obtained by multiple antenna arrangements in the form of array beams to overcome the loss of millimeter waves, and the short wavelength of millimeter waves also makes massive MIMO multi-antenna become a reality. Therefore, millimeter-wave massive MIMO communication has become a hot issue in recent years, but traditional all-digital precoding is difficult to implement due to high consumption and high hardware cost of the huge number of antennas at the transceiver end. In order to reduce system cost while maintaining good performance, millimeter-wave massive MIMO hybrid precoding technology is proposed [4]. Hybrid precoding consists of a large-scale analog precoder and a small-scale digital precoder. How to make the hybrid performance approach the theoretical optimal all-digital precoding has become a key issue in this field [5-6].

The emergence of hybrid precoding combines a high-dimensional analog precoder with a low-dimensional digital encoder, which effectively reduces the number of radio frequency (RF) links, and greatly reduces system energy consumption and hardware costs. However, analog precoding generally only undertakes the problem of phase transformation and needs to satisfy the constant modulus constraint. Therefore, the approximation of hybrid precoding to the theoretical all-digital precoder is designed as a non-convex optimization problem [7]. Hybrid precoding often fails to achieve optimal system performance due to the limited RF chain. Owing to the inconsistency between the dimensions of the data terminal and the transmitter terminal, the method of matrix singular value decomposition (SVD) can often achieve better results. Reference [8] proposes a method based on Orthogonal Matching Pursuit (OMP) hybrid precoding, using the sparseness of the millimeter wave channel to find the closest solution to the all-digital precoder from the transmitter steering vector as the RF chain, but when the number of RF chains is no enough, it is difficult to support the entire channel space, resulting in accuracy is not high. Reference [9] proposed a phase extraction-based alternating optimal (PE-AltMin) hybrid precoding, through the phase extraction of the analog encoder, combined with the repeated alternation of the digital precoding matrix after matrix decomposition. Therefore, better spectral efficiency is obtained, which is often used as the comparison object of other algorithms, but the complexity is extremely high due to the need for multiple iterations. References [10-11] use geometric mean decomposition (GMD) on the basis of SVD decomposition to make the signal with the lowest energy cross the noise level by geometric mean. This method has good performance, but the calculation is relatively complicated. References [12-14] take the analog encoder and the channel as a joint one, and then iterate repeatedly through SVD. Although this algorithm has good design performance, it only interates the analong encoder.So that there is considerable room for improvement in spectral efficiency.

The main contributions of this paper are as follows: (1) The pseudo-inverse iterative functions of the analog coding matrix and the digital coding matrix are derived under the condition that the digital coding matrix is not square. It avoids the problem of increasing the complexity caused by the need to supplement the support vector when the number of data chains and the number of radio frequency chains are not equal; (2) The concept of digital-analog union is proposed. The joint iteration of analog precoding and digital precoding matrix enhances the overall connection between the digital-to-analog coding matrix and the channel matrix, and reduces the information loss caused by the quantization error caused by digital-to-analog conversion during the system transmission process. (3) A millimeter-wave massive MIMO precoding simulation system is established. Experiments on various antenna systems show that the algorithm proposed in this paper not only improves the spectral efficiency, but also ensures the stability, and has good results in various antenna systems.

We use the following notation throughout this paper: $\mathbf{A}$ is a matrix; $\mathbf{a}$ is a vector; a is a scalar; $\mathbf{A}^{(i)}$, $\mathbf{A}$(m,n) and angle($\mathbf{A}$) represent the i^{th}column of $\mathbf{A}$, the $(m,n)^{th}$ element of $\mathbf{A}$, and a matrix containing the phases of the entries of $\mathbf{A}$ respectively; $(\cdot)^T$ and $(\cdot)^H$ denote transpose and conjugate transpose respectively; $\|\mathbf{A}\|_F$ is the Frobenius norm of $\mathbf{A}$, tr($\mathbf{A}$) is its trace and $|\mathbf{A}|$ is its determinant; diag($\mathbf{A}$) is a vector formed by diagonal elements of $\mathbf{A}$; $\mathbf{I}_N$ is the N×N identify matrix; $\mathcal{CN}(\mathbf{a};\mathbf{A})$ is a comlex Gaussian vector with mean a and covariance matrix $\mathbf{A}$. Expecation is denoted by $\mathbb{E}[\cdot]$

2 System Model

The typical single-user mmWave massive MIMO system model in the field used in this paper is shown in Fig1. Base Station (BS), in which a transmitter with N_t antennas communicates N_s date streams to receiver with N_r antennas.The transmitter and receiver are respectively deployed N_t^{RF} and N_r^{RF} data chains, such that $N_s \leq N_t^{RF} \leq N_t$, $N_s \leq N_r^{RF} \leq N_r$, $N_t^{RF} = N_r^{RF} =$

N_{RF}. Figure 1 shows the complete model of hybrid precoding for mmWave massive MIMO system. Considering the results of the combined action of the precoder and the combiner, according to the above system model, the received signal is:

$$\begin{aligned} \mathrm{y} &= \sqrt{\rho}\mathrm{W}^H\mathrm{HFs} + \mathrm{W}^H\mathrm{n} \\ &= \sqrt{\rho}\mathrm{W}_{\mathrm{BB}}^H\mathrm{W}_{\mathrm{RF}}^H\mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{BB}}\mathrm{s} + \mathrm{W}_{\mathrm{BB}}^H\mathrm{W}_{\mathrm{RF}}^H\mathrm{n} \end{aligned} \tag{1}$$

where y is the $N_r \times 1$ and s is the $N_s\times$ 1 signal vector, such that $\mathbb{E}[ss^H]$ $= \frac{1}{N_s}\mathbf{I}_{N_s}$. $\mathbf{H}$ is the $N_r \times N_t$ changnl matrix. ρ represents the average received power, and $\mathbf{n}$ is the vector of i.i.d $\mathcal{CN}(0, \sigma_n^2)$ noise. W_{RF} is the $N_r \times N_{RF}$ RF combining matrix and W_{BB} is the $N_{RF} \times N_s$ baseband combining matrix using at the receiver. Similar to that, F_{RF} and F_{BB}, represent $N_t \times N_{RF}$ RF precoding and $N_{RF} \times N_s$ baseband precoding matrix at trassmitter represently.

Since the analog precoder(combiner) uses a phase shifter to connect each RF chain with all antennas, constant mode constraints need to be met:

$$|\mathrm{F}_{\mathrm{RF}}(\mathrm{i},\mathrm{j})| = 1, |\mathrm{W}_{\mathrm{RF}}(\mathrm{i},\mathrm{j})| = \frac{1}{N_s} \tag{2}$$

And the digital precoder(combiner) and analog precoder(combiner) matrices need to satisfy the total power limit:

$$\|\mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{BB}}\|_{\mathrm{F}}^2 = N_s, \|\mathrm{W}_{\mathrm{RF}}\mathrm{W}_{\mathrm{BB}}\|_{\mathrm{F}}^2 = N_s \tag{3}$$

Assuming that the channel state CSI is completely known to the transceiver, according to [15], the following formula is generally used as the upper limit of the system spectral efficiency:

$$\begin{aligned} R = \log_2(|\mathrm{I}_{N_s} + \frac{\sqrt{\rho}}{N_{\mathrm{s}}}\mathrm{R}_{\mathrm{n}}^{-1}\mathrm{W}_{\mathrm{BB}}^H\mathrm{W}_{\mathrm{RF}}^H\mathrm{H}\mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{BB}} \\ \times\mathrm{F}_{\mathrm{BB}}^H\mathrm{F}_{\mathrm{RF}}^H\mathrm{H}^H\mathrm{W}_{\mathrm{RF}}\mathrm{W}_{\mathrm{BB}})| \end{aligned} \tag{4}$$

where $\mathrm{R_n}$=$\sigma_{\mathrm{n}}^2\mathrm{W}_{\mathrm{BB}}^H\mathrm{W}_{\mathrm{RF}}^H\mathrm{W}_{\mathrm{RF}}\mathrm{W}_{\mathrm{BB}}$ is the noise covariance matrix after combining.

3 Channel Model

Due to the large number of antennas and the sparse nature of mmWave massive MIMO channels, traditional channels are no longer applicable.This paper adopts the popular Saleh-Valenzuela millimeter-wave channel model in the industry [15].Suppose the number of scattering clusters is N_{cl}, Each cluster contains N_ray transmission paths, The channel matrix $\mathbf{H}$ can be expressed as:

$$H = \gamma \sum_{i}^{N_{cl}} \sum_{i}^{N_{ray}} \alpha_{i,j} a_r(\phi_{i,j}^r, \theta_{i,j}^r) a_t(\phi_{i,j}^t, \theta_{i,j}^t) \tag{5}$$

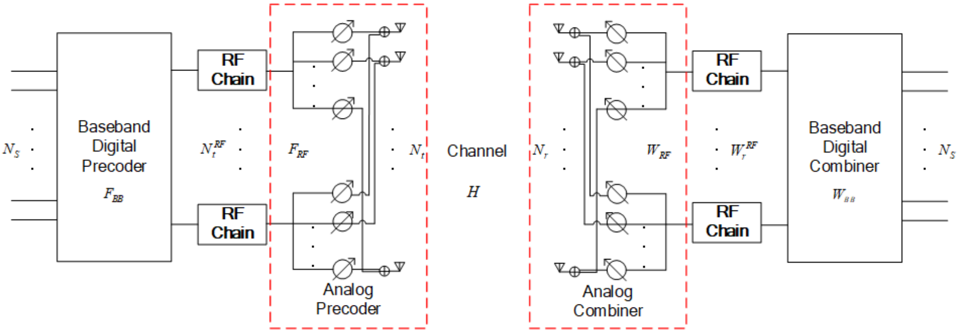

Fig. 1 The block diagram of the hybrid structure of single-user transceiver in mmwave MIMO system.

where γ is a normalization factor such that $\gamma = \sqrt{\frac{N_t N_r}{N_{cl} N_{ral}}}$, and $\alpha_{i,j}$ is the complex gain of the j^{th} ray in the i^{th} scattering cluster following the complex Gaussian distribution $\mathcal{CN}(0;\sigma^2_{\alpha,i})$. We assume that $\sum_{i=1}^{N_{cl}} \sigma^2_{\alpha,i} = \gamma$, so that $\mathbb{E}[\|\mathbf{H}\|_F^2] = N_t N_r$. $\mathbf{a}_r(\phi^r_{i,j}, \theta^r_{i,j})$ and $\mathbf{a}_t(\phi^t_{i,j}, \theta^t_{i,j})$ represent the corresponding steering vectors of the antenna array at the transceiver end, respectively; $\theta^r_{i,j} \in [0, 2\pi)$ and $\theta^r_{i,j} \in [0, 2\pi)$ represent the horizontal arrival angle and departure angle of the j^{th} transmission path of the i^{th} cluster, respectively.Similarly, $\phi^r_{i,j} \in [0, 2\pi)$ and $\phi^r_{i,j} \in [0, 2\pi)$ represent the vertical arrival angle and departure angle of the j^{th} transmission path of the i^{th} cluster. respectively.

Assuming that the BS transceivers are deployed in a uniform plane array (UPA), the steering vector of the transceivers is:

$$\mathbf{a}_{\text{UPA}}(\phi, \theta) = \frac{1}{\sqrt{N_t}}[1, ..., e^{j\frac{2\pi}{\lambda}d(msin(\phi)sin(\theta)+ncos(\theta))}, \\ , ..., e^{j\frac{2\pi}{\lambda}d((W-1)msin(\phi)sin(\theta)+(H-1)ncos(\theta))}]^T \tag{6}$$

where N represents the number of antenna elements; λ represents the millimeter wave wavelength, d represents the distance between antenna elements, $0 \leq m \leq W$, $0 \leq n \leq H$ are the y and z indices of an antenna element respectively and the antenna array size satisfy $N = WH$.

4 DJEC-HP Algorithm

The DJEC-HP algorithm proposed in this paper first combines the analog encoder and the channel into an equivalent channel $\mathrm{H_e}$. Digital encoder is then obtained through SVD decomposition, and finally the analog encoder is iteratively regenerated through the pseudo-inverse equation derived above.The purpose of the joint channel is to better eliminate the information loss ,that caused by the quantization error of digital-to-analog conversion in the system propagation process of the data link. Strengthening the connection among the channel, the analog encoder and the digital encoder, so that the obtained hybrid precoding design is closer to the theoretical optimal digital precoder.In this paper, the design goal is to optimize spectrum efficiency, let the designed system approaches the theoretical upper limit of spectrum:

$$\begin{aligned}
\{\mathrm{F_{RF}}, \mathrm{F_{BB}}, \mathrm{W_{RF}}, \mathrm{W_{BB}}\} = & \underset{\mathrm{F_{RF}},\mathrm{F_{BB}},\mathrm{W_{RF}},\mathrm{W_{BB}}}{\arg\max} \\
& \log_2(|\mathrm{I}_{N_\mathrm{s}} + \frac{\sqrt{\rho}}{N_\mathrm{s}}\mathrm{R_n^{-1}}\mathrm{W_{BB}^{\mathit{H}}}\mathrm{W_{RF}^{\mathit{H}}}\mathrm{H}\mathrm{F_{RF}}\mathrm{F_{BB}} \\
& \times\mathrm{F_{BB}^{\mathit{H}}}\mathrm{F_{RF}^{\mathit{H}}}\mathrm{H}^{H}\mathrm{W_{RF}}\mathrm{W_{BB}}|) \\
s.t. \quad & |\mathrm{F_{RF}}(\mathrm{i},\mathrm{j})| = 1, \forall \mathrm{i},\mathrm{j} \\
& |\mathrm{W_{RF}}(\mathrm{i},\mathrm{j})| = 1, \forall \mathrm{i},\mathrm{j} \\
& |\mathrm{F_{RF}}\mathrm{F_{RF}}|_\mathrm{F}^2 = \mathrm{N_s}
\end{aligned} \tag{7}$$

4.1 Joint Channel Design Theory

First, consider combining the RF part and the channel into a whole, and then perform matrix decomposition on this joint whole to obtain digital encoders $\mathrm{F_{BB}}$ and $\mathrm{W_{BB}}$. This joint design idea aims to reduce the information loss caused by dimensional transformation during digital-to-analog conversion.The equivalent channel in this paper is inspired by the literature [5]:

$$\mathrm{H_e} = \mathrm{W_{RF}^{\mathit{H}}}\mathrm{H}\mathrm{F_{RF}} \tag{8}$$

The SVD decomposition of $\mathrm{H_e}$ is as follows:

$$[\mathrm{U_e}\Sigma_\mathrm{e}\mathrm{V_e}] = \mathrm{SVD}(\mathrm{H_e}) \tag{9}$$

At this time, it is obvious that the digital coding matrix $\mathrm{F_{BB}}$ and $\mathrm{W_{BB}}$ is easy to obtain:

$$\mathrm{F_{BB}} = \mathrm{V_e}[:, 1 : N_s] \tag{10}$$

$$\mathrm{W_{BB}} = \mathrm{U_e}[:, 1 : N_s] \tag{11}$$

The digital coding matrix here, in the massive MIMO system, the digital chain propagation process is carried out independently and in parallel without interfering with each other.Therefore, after implementing some measures, it can be assumed that $\mathrm{F_{BB}}$ and $\mathrm{W_{BB}}$ are both unitary matrices, so that $\mathrm{F_{BB}^{\mathit{H}}}\mathrm{F_{BB}} = \mathrm{I}_{N_s}$,$\mathrm{W_{BB}^{\mathit{H}}}\mathrm{W_{BB}} = \mathrm{I}_{N_s}$.

Thus we can derive:

$$\mathrm{V_e^{\mathit{H}}}\mathrm{F_{BB}}\mathrm{F_{BB}^{\mathit{H}}}\mathrm{V_e} = \begin{bmatrix} \mathrm{I}_{N_s} & 0 \\ 0 & 0 \end{bmatrix} \tag{12}$$

Define a diagonal matrix $\tilde{\Sigma}$ whose elements are the squares of eigenvalues of N_s items of $\mathrm{H_e}$:

$$\tilde{\Sigma} = \begin{bmatrix} \lambda_1^2 & \cdots & 0 \\ \vdots & \ddots & 0 \\ 0 & \cdots & \lambda_{N_s}^2 \end{bmatrix} \tag{13}$$

Therefore, we can obtain the following simplification:

$$\mathrm{W}_{\mathrm{BB}}^{H}\mathrm{H}\mathrm{W}_{\mathrm{RF}}^{H}\mathrm{H}\mathrm{F}_{\mathrm{RF}}^{H}\mathrm{F}_{\mathrm{BB}}\mathrm{F}_{\mathrm{BB}}^{H}\mathrm{F}_{\mathrm{RF}}^{H}\mathrm{H}\mathrm{W}_{\mathrm{RF}}\mathrm{W}_{\mathrm{BB}} = \tilde{\Sigma} \tag{14}$$

According to literature [8], it is found that in massive MIMO systems, the analog combiner approximates the unitary matrix:

$$\mathrm{W}_{\mathrm{RF}}^{H}\mathrm{W}_{\mathrm{RF}} \approx N_r \mathrm{I}_{N_{RF}} \tag{15}$$

Then we can get a further simplification:

$$\mathrm{R_n} \approx N_r \sigma_n^2 \mathrm{W}_{\mathrm{BB}}^{H} \mathrm{I}_{N_{RF}} W_{BB} = N_r \sigma_n^2 \mathrm{I}_{N_s} \tag{16}$$

In summary,let $\beta = \frac{\rho}{N_s N_r \sigma_n^2}$ the upper limit of the system spectrum of formula (4) can be finally simplified as:

$$\begin{aligned} R &= \log_2(|\mathrm{I}_{N_s} + \ \mathrm{W}_{\mathrm{BB}}^{\mathrm{H}}\mathrm{W}_{\mathrm{RF}}^{\mathrm{H}}\mathrm{H}\mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{BB}}\mathrm{F}_{\mathrm{BB}}^{\mathrm{H}}\mathrm{F}_{\mathrm{RF}}^{\mathrm{H}}\mathrm{H}^{\mathrm{H}}\mathrm{W}_{\mathrm{RF}}\mathrm{W}_{\mathrm{BB}}|) \\ &\approx \log_2(|I_{N_s} + \beta\tilde{\Sigma}|) = \sum_{i=1}^{N_s} \log_2(1 + \beta\lambda_i^2) \end{aligned} \tag{17}$$

So far, the maximum spectrum of the system equivalent to make the square of the eigenvalue of the first N_s term of $\mathrm{H_e}\mathrm{H_e^H}$ maximum, that is, the trace maximum:

$$\begin{aligned} \mathrm{tr}(\mathrm{H_e}\mathrm{H_e}^{H}) &= \mathrm{tr}(\mathrm{W}_{\mathrm{RF}}^{H}\mathrm{H}\mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{RF}}^{H}\mathrm{H}^{H}\mathrm{W}_{\mathrm{RF}}) \\ &= \|\mathrm{W}_{\mathrm{RF}}^{H}\mathrm{H}\mathrm{F}_{\mathrm{RF}}\|_{\mathrm{F}}^2 \end{aligned} \tag{18}$$

Ultimately, the question turns into how to design $\mathrm{F_{RF}}$ and $\mathrm{W_{RF}}$ to be maximum, namely:

$$\{\mathrm{F_{RF}}, \mathrm{W_{RF}}\} = \arg\max_{\mathrm{F_{RF}},\mathrm{W_{RF}}} \|\mathrm{W}_{\mathrm{RF}}^{H}\mathrm{H}\mathrm{F}_{\mathrm{RF}}\|_{\mathrm{F}}^2 \tag{19}$$

According to the literature [8], the non-convex optimal hybrid precoding design problem is approximately transformed into the minimum Frobenius norm distance problem (20). Due to the similarity in the structure of the transceiver, we take the transmitter as an example to design:

$$\{\mathrm{F_{RF}}, \mathrm{F_{BB}}\} = \arg\min_{\mathrm{F_{RF}}} \|\mathrm{F_{opt}} - \mathrm{F_{RF}}\mathrm{F_{BB}}\|_{\mathrm{F}}^2 \tag{20}$$

It is known from (10) that F_{BB} can be generated by the joint channel. Inspired by the literature [17], F_{BB} is fixed and then expanded:

$$\begin{aligned}\{\mathrm{F}_{\mathrm{RF}}\} &= \underset{\mathrm{F}_{\mathrm{RF}}}{\arg\min}\|\mathrm{F}_{\mathrm{opt}} - \mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{BB}}\|_{\mathrm{F}}^2 \\ &= N_s - 2\mathrm{tr}(\mathrm{F}_{\mathrm{opt}}^H\mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{BB}}) + \mathrm{tr}(\mathrm{F}_{\mathrm{BB}}^H\mathrm{F}_{\mathrm{RF}}^H\mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{BB}})\end{aligned} \tag{21}$$

Taking the derivative of the above formula (21) with respect to F_{RF}, after making it zero, we get:

$$\mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{BB}}\mathrm{F}_{\mathrm{BB}}^H = \mathrm{F}_{\mathrm{opt}}\mathrm{F}_{\mathrm{BB}}^H \tag{22}$$

Considering that the number of N_s is not equal to the number of N_{RF} chains in general (A<B), using the pseudo-inverse solution can avoid errors in the matrix dimension:

$$\tilde{\mathrm{F}}_{\mathrm{RF}} = \mathrm{F}_{\mathrm{opt}}\mathrm{F}_{\mathrm{BB}}^H(\mathrm{F}_{\mathrm{BB}}\mathrm{F}_{\mathrm{BB}}^H)^\dagger \tag{23}$$

So far, the iterative solutions related to F_{RF} and F_{BB} have been obtained. In order to ensure the constant modulus constraint, the phase is taken as:

Algorithm 1 Digital Joint Equivalent Channel (DJEC) Hybrid precoding.

Input:H
1:Performance SVD decomposition of H, $\mathrm{F}_{\mathrm{opt}}$ and $\mathrm{W}_{\mathrm{opt}}$ are obtained;
2:Construct $\mathrm{F}_{\mathrm{RF}}^{(0)}$ and $\mathrm{W}_{\mathrm{RF}}^{(0)}$ with random phases and set i=0
3: for i = 1:K
4: $\mathrm{H}_{\mathrm{e}} = \mathrm{W}_{\mathrm{RF}}^H\mathrm{H}\mathrm{F}_{\mathrm{RF}}$;
5: $[\mathrm{U}_{\mathrm{e}}\Sigma_{\mathrm{e}}\mathrm{V}_{\mathrm{e}}] = \mathrm{SVD}(\mathrm{H}_{\mathrm{e}})$;
6: $\mathrm{F}_{\mathrm{BB}} = \mathrm{V}_{\mathrm{e}}[:, 1 : N_s], \mathrm{W}_{\mathrm{BB}} = \mathrm{U}_{\mathrm{e}}[:, 1 : N_s]$;
7: $\tilde{\mathrm{F}}_{\mathrm{RF}} = \mathrm{F}_{\mathrm{opt}}\mathrm{F}_{\mathrm{BB}}^H(\mathrm{F}_{\mathrm{BB}}\mathrm{F}_{\mathrm{BB}}^H)^\dagger$;
8: $\tilde{\mathrm{W}}_{\mathrm{RF}} = \mathrm{W}_{\mathrm{opt}}\mathrm{W}_{\mathrm{BB}}^H(\mathrm{W}_{\mathrm{BB}}\mathrm{W}_{\mathrm{BB}}^H)^\dagger$;
9: $\mathrm{F}_{\mathrm{RF}} = \mathrm{e}^{\mathrm{j}\varphi(\tilde{\mathrm{F}}_{\mathrm{RF}})}, \mathrm{W}_{\mathrm{RF}} = \mathrm{e}^{\mathrm{j}\varphi(\tilde{\mathrm{W}}_{\mathrm{RF}})}$;
10: end
11:$\mathrm{F}_{\mathrm{BB}} = \frac{\sqrt{N_s}}{\|\mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{BB}}\|_{\mathrm{F}}}\mathrm{F}_{\mathrm{BB}}$, $\mathrm{W}_{\mathrm{BB}} = \frac{\sqrt{N_s}}{\|\mathrm{W}_{\mathrm{RF}}\mathrm{W}_{\mathrm{BB}}\|_{\mathrm{F}}}\mathrm{W}_{\mathrm{BB}}$
Output:$\mathrm{F}_{\mathrm{RF}}, \mathrm{F}_{\mathrm{BB}}, \mathrm{W}_{\mathrm{RF}}, \mathrm{W}_{\mathrm{BB}}$

where K is the number of iterations, as discussed below.It can be seen from the above algorithm flow that it can be roughly divided into two steps. First, the analog coding matrix F_{RF}, W_{RF} and the channel are combined into a whole.Then, the digital coding matrices F_{BB} and W_{BB} are obtained through matrix decomposition, and then the analog coding matrix F_{RF} and W_{RF} is regenerated by formula (23).The hybrid precoding matrix is obtained by this iterative method that $\mathrm{F}_{\mathrm{hybrid}} \approx \mathrm{F}_{\mathrm{RF}}\mathrm{F}_{\mathrm{BB}}$ and $\mathrm{W}_{\mathrm{hybrid}} \approx \mathrm{W}_{\mathrm{RF}}\mathrm{W}_{\mathrm{BB}}$.

The time complexity comparison of OMP sparse precoding [8], joint channel precoding [5], PE-AltMin precoding [9] and the AI-JCHP algorithm proposed in this paper is shown in Table 1:

Table 1 Comparison of time complexity of different algorithms.

Algorithms	**Time Complexity**
OMP Hybrid Precoding[8]	$O(N_t^3 + 2N_{cl}N_{ray}N_{RF}N_t^2)$
Joint Hybrid Precoding[5]	$O(24N_{RF}^2 N_t^2)$
PE-AltMin Hybrid Precoding[9]	$O[2\alpha(N_t^2 N_{RF}^2 + N_{RF}^3 + N_t^4)]$
Proposed Algorithms	$O[K(8N_{RF}^2 N_t^2 + N_{RF}^3)]$

where Table 1 is a rough calculation of the main complexity parts (SVD decomposition, data iteration) of different algorithms under the premise of time complexity $O(N_t){\approx}O(N_r)$ and $O(N_s){\approx}O(N_{RF})$.In PE-AltMin hybrid precoding, α represents the number of iterations. We use the error between the design hybrid coding and the theoretical all-digital coding as the threshold to stop the iteration. Therefore, although the algorithm has good design performance, it is extremely complex and difficult to implement. K in this proposed algorithm represents the number of iterations. In order to compare with the algorithm in the literature [5], the same iteration number of 8 is adopted.From the simulation analysis, although OMP hybrid precoding has the lowest complexity (introducing complexity $O(N_t^3 + 2N_{cl}N_{ray}N_{RF}N_t^2)$), it has the worst performance.On the contrary, PE-AltMin has enough excellent performance, but it leads to a significant increase in complexity (introduction of complexity $O[2\alpha(N_t^2 N_{RF}^2 + N_{RF}^3 + N_t^4)]$).Therefore, how to find their own balance between complexity and performance is another difficulty in this field. Joint precoding and this proposed algorithm combine the analog coding matrix with the channel, which reduces the complexity (introducing $O(24N_{RF}^2 N_t^2)$ and $O[K(8N_{RF}^2 N_t^2 + N_{RF}^3)]$ complexity respectively), but has good performance.

5 Simulation Analysis

To verify the feasibility and effectiveness of the proposed algorithm, the mmWave simulation environment adopts the industry-wide Saleh-Valenzuela channel model.Assume that the propagation environment is cluster $N_{cl} = 8$, and each cluster has $N_{ray} = 10$ paths.For simplicity, it is assumed that each cluster is of equal energy, satisfying $\sigma_{\alpha,i} = \sigma_\alpha, \forall i$. And it is assumed that the angular propagation of the transmitter and the receiver is equal in azimuth and elevation, satisfying $\sigma_\phi^t = \sigma_\phi^r = \sigma_\theta^t = \sigma_\theta^r$, see Table 2 for details:

Literature [18] found that when $2N_s{\le}N_{RF}$, the performance theory of hybrid precoding can be close enough to the performance of all-digital precoding, so this paper only considers the scenario that $N_s{\le}N_{RF}{\le}2N_s$.When it is outdoors, the arrangement of reducing interference and increasing beamforming gain should be adopted. The directional antenna element array in literature

Table 2 Simulation Parameters.

Parameter Name	Parameter Value
Number of Transmitter Antennas N_t	128∼256
Number of Receiver Antennas N_r	16∼64
Number of RF chains at the Transmitter N_{RF}^t	2∼4
Number of RF chains at the Receiver N_{RF}^r	2∼4
Number of Data Streams N_s	2
Number of Channel Clusters N_{cl}	8
Number of Paths in Each Cluster N_{ray}	10
Array Antenna Type	UPL
Spacing of Antenna Elements d	half wavelength 0.5λ
AOD	[-30,30]
EOD	[80,100]
AOA	[180,180]
EOA	[0,180]
Channel Distribution	Laplace Distribution
Simulation Times	1000

[8] is used, and the angular sector of the transmitter is 60° (AOD) and the elevation angle is 20° (EOD).Because the position and orientation of receivers in the actual system are random, they have relatively small omnidirectional element antenna arrays (AOA, EOA), which can receive signals in any direction, and the channel obeys the Laplace distribution.

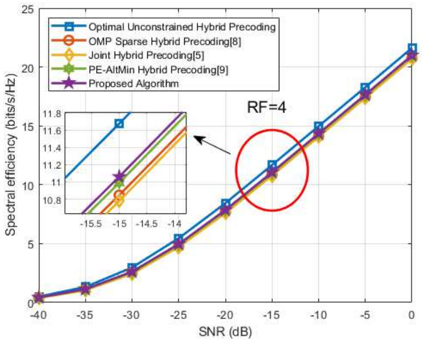

Fig. 2 Spectral efficiency of different algorithms for 256×64 (RF=4) antenna system.

Fig 2 and 3 are comparative simulations under the condition of changing the number N_{RF} of radio frequency chains under the 256×64 antenna system. When the number of RF chains N_{RF} is reduced from 3 to 2, we can clearly observe that with the reduction of the number of RF chains N_{RF}, the design effects of other algorithms have changed.The most obvious changes are OMP hybrid precoding and joint hybrid precoding.The joint hybrid precoding only considers the iteration of the analog encoder and ignores the connection with

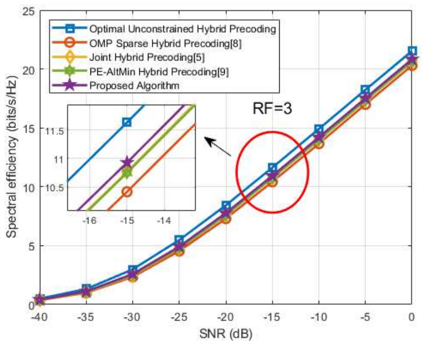

Fig. 3 Spectral efficiency of different algorithms for 256×64 (RF=3) antenna system.

the digital encoder in the design. The OMP hybrid precoding replaces the radio frequency space basis with the steering vector of the maximum linear correlation in the design.However, when the number N_{RF} of radio frequency chains is insufficient, the alternative steering vector cannot support the entire radio frequency sparse space.Therefore, in theory, the two algorithms are seriously affected by the number N_{RF} of radio frequency chains, and the simulation results can also effectively prove this point.This algorithm is optimized on the basis of joint channel hybrid precoding, introduces the idea of iteration, makes full use of the relationship between analog precoding and digital precoding, and has better stability on the premise that the spectral efficiency is higher than the above algorithm.

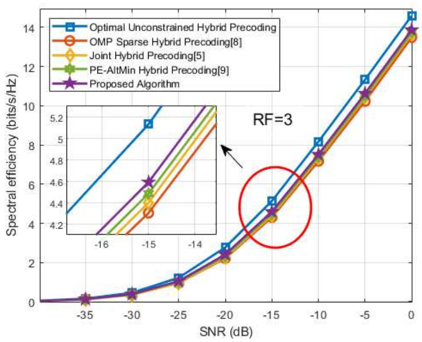

Fig. 4 Spectral efficiency of different algorithms for 64×16 (RF=3) antenna system.

Fig.4 and Fig.5 respectively show the spectral efficiency performance comparison of the above algorithms in the scenario of a 64×16 antenna system. From the fig5, we can clearly see that the proposed algorithm still has better stability than other algorithms in a system with a small number of antennas (64×16). Its performance is basically in the optimal position, and will not fluctuate violently with the change of the radio frequency chain N_{RF}.

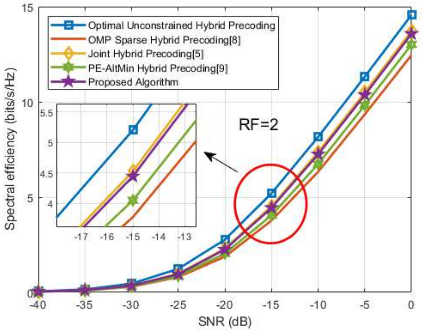

Fig. 5 Spectral efficiency of different algorithms for 64×16 (RF=3) antenna system.

Combining Figures 2, 3, 4, and 5, it can be seen that the algorithm proposed in this paper has better stability than other hybrid precoding algorithms while improving the spectral efficiency. Regardless of the change in the number of antennas or the number N_{RF} of radio frequency chains, the performance of the algorithm basically maintains a good design performance, and it is universally applicable to various antenna systems.

Table 3 Spectral efficiency of each antenna system under -15dB SNR.

Antenna Systems (Class 4 Systems)	**Algorithm** (5 Types of Algorithms)	**Spectral Efficiency** (bit/s/Hz)
256×64(RF=4)	Optimal Precoding	11.67
	OMP Hybrid Precoding	10.85
	Joint Hybrid Precoding	10.78
	PE-AltMin Precoding	10.99
	Proposed Algorithm	11.06
256×64(RF=3)	Optimal Precoding	11.63
	OMP Hybrid Precodingr	10.42
	Joint Hybrid Precoding	10.77
	PE-AltMin Precoding	10.75
	Proposed Algorithm	10.93
64×16(RF=3)	Optimal Precoding	5.14
	OMP Hybrid Precoding	4.31
	Joint Hybrid Precoding	4.40
	PE-AltMin Precoding	4.49
	Proposed Algorithm	4.59
64×16(RF=2)	Optimal Precoding	5.20
	OMP Hybrid Precoding	3.77
	Joint Hybrid Precoding	4.52
	PE-AltMin Precoding	4.05
	Proposed Algorithm	4.44

6 Conclution

Hybrid precoding aims to improve the spectral efficiency of large-scale mmWave systems while reducing energy consumption and hardware costs, which is of practical significance.In this paper, an iterative hybrid precoding algorithm based on joint channel is proposed. First, the analog coding matrix and the channel matrix are combined into an equivalent matrix $\mathrm{H_e}$, then the digital coding matrix is obtained by SVD decomposition, and finally the digital coding matrix is combined with the optimal all-digital coding to regenerate the analog precoding matrix. Compared with the joint hybrid precoding matrix before the improvement, this algorithm introduces the iteration of the digital coding matrix, and pays more attention to the overall connection of the coding matrix.The algorithm in this paper integrates the analog coding matrix, the channel matrix and the digital coding matrix, thereby reducing the information loss caused by the dimensional transformation of the digital-analog matrix during the propagation of the data stream in the mmWave system.

Simulation experiments show that the proposed algorithm has higher spectral efficiency than other hybrid precoding when the number of antennas and radio frequency chain changes, and the stability is also ideal, it is not easy to fluctuate, and it is closer to the optimal all-digital precoding.

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