Partially connected hybrid precoding design for millimeter wave MIMO systems

Xinglin Zheng\textsuperscript{1,2*}, Yue Wang\textsuperscript{1,2} and Wei Liu\textsuperscript{1,2}

\textsuperscript{1}Communications Core Chip, Protocol and System Application Innovation Team, Chongqing University of Posts and Telecommunications, Chongqing, 400065, China
\textsuperscript{2}School of Communication and Information Engineering, Chongqing University of Posts and Telecommunications, Chongqing, 400065, China
*Corresponding author’s e-mail: s170131231@stu.cqupt.edu.cn

Abstract. To solve the problem of spectral efficiency and energy efficiency of hybrid precoding in millimeter-wave communication system. This paper considers alternately using the minimization framework to increase beamforming gain. However, such framework because the radio frequency domain phase modulation network are using phase shifters, which result in non-convex constraints, and known solutions suffer from high computational complexity. Before using the alternate minimization architecture, it is proved that the phase angle of the ordered right singular vector of the channel matrix can be used to initialize the analog precoder, and hence, the complicated optimization procedures used to search the near-optimum analog precoding matrix can be avoided. In addition, the antenna array response vectors at the transmitter are not required. The simulation results show that the performance of the proposed algorithm is better than the traditional partial-connected algorithm and the complexity is lower, especially in the case of high signal-to-noise ratio.

1. Introduction
Millimeter-wave multi-input-multiple-output (MIMO) systems are key technology for next-generation communication systems due to their high bandwidth and high spectral efficiency\cite{1-3}. For example, the wavelength is 10 mm at a carrier frequency of 30 GHz, which allows a large number of antennas to be packaged in a small area. Large-scale antenna arrays can provide higher beamforming gain to compensate for the high path loss of millimeter-wave signals\cite{4}. However, the transmission precoding usually only considers the digital domain in the conventional MIMO system, and needs to connect the same number of radio frequency (RF) links as the number of antennas\cite{5-6}. The traditional full-digital precoding is not feasible in millimeter-wave communication systems due to the high hardware cost and power consumption. So the millimeter-wave system combines a large number of antenna elements to offset the path and penetration loss. In order to solve this problem, the concept of hybrid precoding was proposed in\cite{7}, replacing the traditional full-digital precoder with a two-stage precoder. The signal is first processed with a small digital precoder and then with a large analog precoder to obtain a high beamforming gain.

At present, the hybrid precoding scheme has a Fully-Connected (FC) and Partially Connected (PC) structure\cite{8}, and the transmitting end system model is shown in figure 1. Each RF chain is connected to all antennas through a large number of phase shifters to maximize the achievable transmission gain in a FC structure, but will result in high power consumption\cite{9}. In contrast, a smaller number of phase
shifters is required in PC structures, which can lead to higher energy efficiency although loss of performance[10]. For the study of FC hybrid precoding, the work in[11] proposed an algorithm based on the manifold optimization alternating minimization algorithm (MO-Alt) to approximate the performance of the full-digital precoder. However, the analog precoder was updated by linear search algorithms, and nested loops slowed down the entire solution process with high complexity. In[12], the DFT codebook was used to form the analog precoding matrix, which greatly reduces the complexity of the algorithm. However, there was no good matching channel and the performance gain was small since the analog precoding was selected from the DFT codebook set. For the PC hybrid precoding due to its lower energy consumption, it is a research interest in recent years. Yu et al.[11] used the idea of semidefinite relaxation (SDR), an alternating minimization algorithm was proposed. The digital precoder was solved by the convex optimization toolbox (CVX), which each iteration needed standard convex optimization. The algorithm solveed the problem of semidefinite programming (SDP), resulting in high complexity. In[13], the water-filling algorithm was used to design digital precoding, and the analog precoding design was equivalent to the single stream optimal transmitter beamforming problem under single antenna power constraints, which had high complexity.

![Figure 1. Transmitting hybrid precoding system model: (a) PC structure; (b) FC structure](image_url)

This paper studies PC hybrid precoding considering the practical application of the project. The phase angle of the channel right singular value vector is proved to initialize the feasibility of the analog precoder. In order to obtain better performance and higher energy efficiency, an alternate iteration singular value decomposition (Alt-ISVD) hybrid precoding algorithm is designed. The simulation results show that the proposed algorithm is better than the traditional PC algorithm with the increase of signal-to-noise ratio, and the complexity is significantly lower than that of the proposed algorithm in[13]. Compared with the traditional orthogonal matching pursuit (OMP) algorithm[7] has higher energy efficiency.

2. System model and channel model

2.1. System model

Considering the downlink millimeter-wave single-user system, the base station (BS) side is equipped with \(N_t\) transmit antennas, and the number of transmit-end RF links is \(N_{RF}^t\). The system model is shown in figure 1(a). Each RF link is connected to only one independent sub-array, and each independent sub-array contains \(M\) antennas, ie \(M = N_t / N_{RF}^t\). It can be seen from Figure 1 that when the BS is equipped with \(N_t\) antennas, the analog domain of FC requires \(N_t N_{RF}^t\) phase shifters to be connected to all antennas, while the PC analog domain requires only \(N_t\) phase shifters to be connected to partial antennas. Therefore, hardware cost and power consumption is lower by using PC structures. The transmitter has \(N_s\) data stream transmission, and the data stream \(s \in \mathbb{C}^{N_s \times 1}\) is first preprocessed in...
the digital domain by the digital baseband precoder $F_{bb} \in \mathbb{C}^{M \times N_t}$, and then modulate into the antenna elements through the analog precoder $F_{rf} \in \mathbb{C}^{N_t \times NF}$ in the analog domain to form a beamforming gain. After a two-stage precoder, the final transmitted signal is

$$x = F_{bb}F_{bb}s$$

(1)

Where the transmitted signal $s$ should satisfy $E[ss^H] = \frac{1}{N_t}I_{N_t}$, and the analog precoder and the digital precoder should satisfy the total power limit $\|F_{bb}F_{bb}\|^2 = N_t$. Where $E[\bullet]$, $I$ and $\|\bullet\|_F$ represents the expectation, identity matrix, and Frobenius norm, respectively. After the receiver is decoded, the signal is

$$y = \sqrt{\rho}W_{bb}^H F_{rf}^H F_{bb}s + W_{bb}^H W_{bb}^H n$$

(2)

Where $\rho$ is the received power and $W_{bb}$ and $W_{rf}$ are represented as the receiver digital combiner and the analog combiner, respectively. $F_{rf}$ and $W_{rf}$ should meet the constant modulus limit, ie $\|F_{rf}\|_F = 1$, $\|W_{rf}\|_F = 1$. It is an $N_t \times N_t$-dimensional millimeter wave geometric channel matrix $H$. $n - \mathcal{C}\mathcal{N}(0, \sigma^2)$ is an additive Gaussian noise obeying the $\mathcal{C}\mathcal{N}(0, \sigma^2)$ distribution. Where $(\bullet)^H$ is the conjugate transpose and $N_t$ is the number of antennas provided by the mobile station (MS).

2.2. Channel model

Considering the limited scattering characteristics of millimeter wave channels, this paper uses the extended Saleh-Valenzuela cluster channel model[7], and the narrow-band millimeter wave channel matrix $H$ is expressed as

$$H = \sqrt{\frac{N_t}{L}} \sum_{i=1}^{L} a_i(\theta) a_i^H(\phi_i)$$

(3)

Where $L$ is the number of system propagation paths and each cluster can generate a propagation path. $a_i$ is the complex gain of the $L$th path obeying the $\mathcal{C}\mathcal{N}(0, \sigma^2)$ distribution, and $a_i(\theta)$ and $a_i(\phi_i)$ are respectively defined as the receive and transmit array vectors, where $\theta \in [0, 2\pi]$ and $\phi_i \in [0, 2\pi]$ are the arrival and departure azimuths of the $L$th path. When the transceiver antenna array is a uniform linear array (ULA), the normalized response vector of the transmit and receive antenna arrays is expressed as follows

$$a_i(\phi_i) = \frac{1}{N_t} \begin{bmatrix} e^{j2\pi d\sin(\phi_i)} & \ldots & e^{j2\pi (N_t-1)d\sin(\phi_i)} \end{bmatrix}^T$$

(4)

$$a_i(\theta) = \frac{1}{N_t} \begin{bmatrix} e^{j2\pi d\sin(\theta)} & \ldots & e^{j2\pi (N_t-1)d\sin(\theta)} \end{bmatrix}^T$$

(5)

Where $\lambda$ is the carrier wavelength and $d$ is the antenna spacing.

2.3. Problem description

The goal of designing hybrid precoding is to reduce the complexity of the algorithm and improve the spectral efficiency and energy efficiency of the system. The system’s spectral efficiency is expressed as

$$R = \log_2\left| I_{N_t} + \frac{\rho}{N_t} R^{-1} W_{bb}^H W_{bb}^H F_{rf} F_{bb} \bullet \right|$$

$$F_{bb}^H F_{bb} H^H W_{bb}^H W_{bb}^H$$

(6)
Where \( \mathbf{R}_n = \sigma^2 \mathbf{W}^\mathsf{H} \mathbf{W} \mathbf{W}^\mathsf{H} \mathbf{W} \mathbf{W} \) represents the noise covariance matrix processed by the receiver. It can be seen from the equation that to maximize the spectral efficiency, it is necessary to jointly optimize \( \mathbf{W}, \mathbf{F}_{\text{RF}}, \mathbf{F}_{\text{BB}}, \mathbf{F}_{\text{RF}}^\mathsf{H}, \mathbf{F}_{\text{BB}}^\mathsf{H} \) four matrices. Joint optimization is a non-convex problem, and In[11], the joint optimization problem in equation (6) is transformed into the independent optimization problem between the transmitter and the receiver by decoupling, and the optimization target of the transmitter is

\[
R = \log_2 \left| I_{\mathbf{n}} + \frac{\rho}{N \sigma_n^2} \mathbf{H} \mathbf{F}_{\text{RF}} \mathbf{F}_{\text{BB}} \mathbf{F}_{\text{RF}}^\mathsf{H} \mathbf{F}_{\text{BB}}^\mathsf{H} \mathbf{H}^\mathsf{H} \right|
\]

(7)

The mathematical formula of the transmitter precode and the receiver combiner optimization problem is similar, except that the precoding matrix design needs to consider the power constraint. Therefore, this paper mainly studies the design of the transmitter precoder, and the maximum spectral efficiency is equivalent to the minimum Euclidean distance between the hybrid precoding matrix and the optimal precoding matrix, ie

\[
\arg\max \mathbf{F}_{\text{RF}} \mathbf{F}_{\text{BB}} = \mathbf{F}_{\text{RF}}^\mathsf{H} \mathbf{F}_{\text{RF}} \mathbf{F}_{\text{BB}} \mathbf{F}_{\text{BB}}^\mathsf{H} \mathbf{H}^\mathsf{H}
\]

(8)

Where \( \mathbf{F}_{\text{opt}} \in \mathbb{C}^{N_t \times N_t} \) is the optimal digital precoder, and SVD is performed on the channel matrix \( \mathbf{H} \), that is, \( \mathbf{H} = \mathbf{U} \Sigma \mathbf{V}^\mathsf{H} \), where \( \mathbf{U} \) and \( \mathbf{V} \) are respectively corresponding left singular value matrix and right singular value matrix, and \( \Sigma \) is a eigenvalue matrix. Taking the first \( N_t \) column of the right singular value matrix as the optimal digital precoder \( \mathbf{F}_{\text{opt}} = \mathbf{F}(\cdot,1:N_t) \). The \( \mathbf{F}_{\text{RF}} \) should satisfy the PC structure form. The block diagonalization matrix is

\[
\mathbf{F}_{\text{RF}} = \begin{bmatrix}
f_1 & 0 & \cdots & 0 \\
0 & f_2 & \cdots & 0 \\
\vdots & \ddots & \ddots & \vdots \\
0 & 0 & \cdots & f_{N_{\text{RF}}}
\end{bmatrix}
\]

(9)

Where \( f_i = [e^{j\theta_i}, e^{j\theta_i}, \ldots, e^{j\theta_i}]^\mathsf{T} \) and \( \theta_i \) is the phase of the \( i \)th phase shifter.

3. Hybrid precoding design

Considering the single-user transmission scenario, this paper uses a PC structure to prove that the angle of ordered right singular value vector of the channel matrix \( \mathbf{H} \) can be used to initialize the analog precoding matrix without the antenna array response vector at the transmitter.

3.1. Initialize the analog precoder design

According to the analysis in[7] and[11], the hybrid precoding optimization problem can be rewritten as

\[
\left( \mathbf{F}_{\text{RF}}^\text{opt}, \mathbf{F}_{\text{BB}}^\text{opt} \right) = \arg\max \mathbf{F}_{\text{RF}}^\mathsf{H} \mathbf{F}_{\text{RF}} \mathbf{F}_{\text{BB}} \mathbf{F}_{\text{BB}}^\mathsf{H} \mathbf{H}^\mathsf{H}
\]

s.t. \( \mathbf{F}_{\text{RF}} \mathbf{F}_{\text{RF}}^\mathsf{H} = 1, \forall i, j \),

\[
\mathbf{F}_{\text{RF}} \mathbf{F}_{\text{BB}} \mathbf{F}_{\text{BB}}^\mathsf{H} = N_t
\]

Minimizing \( \| \mathbf{F}_{\text{opt}} - \mathbf{F}_{\text{RF}} \mathbf{F}_{\text{BB}} \|_F^2 \) and maximizing \( \mathbf{F}_{\text{RF}}^\mathsf{H} \mathbf{F}_{\text{RF}} \mathbf{F}_{\text{BB}} \mathbf{F}_{\text{BB}}^\mathsf{H} \mathbf{H}^\mathsf{H} \) is equivalent[7]. For a given \( \mathbf{F}_{\text{RF}} \), for ease of derivation, use simple least squares to get

\[
\mathbf{F}_{\text{BB}} = \left( \mathbf{F}_{\text{RF}}^\mathsf{H} \mathbf{F}_{\text{RF}} \right)^{-1} \mathbf{F}_{\text{RF}}^\mathsf{H} \mathbf{F}_{\text{BB}}^\text{opt}
\]

(11)
Substituting equation (11) into equation (10)
\[
F_{RF}^{\text{opt}} = \arg \max \text{tr} \left( PF_{\text{opt}} F_{RF}^H \right)
\]
subject to \( \left( F_{RF} \right)_{ij} = 1, \forall i, j, \) (12)

Where, \( \text{tr}(AB) = \text{tr}(BA) \), \( P = F_{RF} \left( F_{RF}^H F_{RF} \right)^{-1} F_{RF}^H \) are applied, \( P \) and \( F_{RF} F_{\text{opt}}^H \) are orthogonal projections. Let \( S_{RF} \) and \( S_{\text{opt}} \) be subspaces of \( F_{RF} \) and \( F_{\text{opt}} \) columns respectively, and \( \text{tr}(PF_{\text{opt}} F_{RF}^H) \) is defined as the inner product of \( S_{RF} \) and \( S_{\text{opt}} \). The problem of \( F_{RF} \) design can be transformed into \( S_{RF} \), and make the inner product of \( S_{RF} \) and \( S_{\text{opt}} \) is the largest, i.e, \( |S_{\text{opt}} \rangle \langle S_{RF}| \) is the largest.

Proposition 1: Given vector \( g = [a, e^{i\phi}, \ldots, a_N e^{i\phi}]^T \), a set \( F = \{ f \in \mathbb{C}^{N+1} : |f(n)| = 1, \forall n \} \), for any \( f \in F \), we have \( f_s = \arg \max |g^H f| = [e^{i\phi}, \ldots, e^{i\phi}]^T \).

Proof: set \( f = [e^{i\phi}, \ldots, e^{i\phi}]^T \), then we have
\[
|g^H f| = \left| \sum_{n=1}^{N} a_n e^{i(n-\delta)} \times \sum_{n=1}^{N} a_n e^{i(\delta-\delta)} \right|^{1/2}
\] (13)

The maximization problem can be converted to take the derivative of \( |g^H f| \) respect to \( \phi \), and make it to equal 0, then we can make \( \phi_i = \phi, i = 1, \ldots, N \). According to Proposition 1, to make the inner product of \( S_{RF} \) and \( S_{\text{opt}} \) the largest, the subspace composed of \( F_{RF} \) and \( F_{\text{opt}} \) is the same, \( F_{\text{opt}} \) is known as the right singular value matrix of the channel matrix, and the corresponding phase angle constitutes its subspace. Due to the unit modulus limitation, \( F_{RF} \) can be formed by the phase angle of the channel right singular value matrix element. From the above analysis, it can be known that the analog precoder can be initialized with the phase angle of the channel matrix right singular value matrix element. The specific initialization process is
\[
H = U \Sigma V^H
\] (14)

Where \( U \) is a \( N_i \times N_i \)-dimensional unitary matrix, \( V \) is an \( N_i \times N_i \)-dimensional unitary matrix, \( \Sigma \) is a diagonal matrix, and diagonal elements are arranged in descending order. Let \( \varphi_{m,n} \) be the phase angle of the \( m \)th row and the \( n \)th column element of \( V \), and take the phase angle of the elements in \( V \) as a set \( F \),
\[
F = \{ v_1, v_2, \ldots, v_{N_i} \}
\] (15)

Where \( v_n = [e^{i\varphi_{m,n}}, e^{i\varphi_{m,n}}, \ldots, e^{i\varphi_{m,n}}]^T \), each of \( v_n \) has the same phase angle as each element corresponding to \( V \). The initialized analog precoding matrix is
\[
F_{RF} = [v_1, v_2, \ldots, v_{N_i}]
\] (16)

That is, the first \( N_{RF} \) column vector of the \( F \) is taken. Applying to (9), performing block diagonalization, the resulting initial analog precoding matrix is
\[
F_{RF} = \begin{bmatrix}
v_1 (1:M) & 0 & \cdots & 0 \\
0 & v_2 (M+1:2M) & \cdots & 0 \\
\vdots & \ddots & \ddots & \vdots \\
0 & 0 & \cdots & v_i ((i-1)M+1:iM)
\end{bmatrix}
\] (17)

Where \( v_i ((i-1)M+1:iM) \) is the \( (i-1)M+1 \)th to \( iM \)th elements in \( v_i \), \( i = 1, 2, \ldots, N_{RF} \), \( M = N_i / N_{RF} \).
3.2. Apply alternating iterative structure design

Applying the alternate minimization structure after obtaining analog precoder $F_{RF}$, fixing the $F_{RF}$, optimize the $F_{BB}$ according to the Euclidean distance of the minimized $F_{opt}$ and $F_{RF}F_{BB}$, and make the

$$F_{BB} = V_i U_i^H$$

(18)

Where, $F_{opt}^H F_{RF} = U_i \Sigma_i V_i^H$, $V_i^k$ is the first $N_i$ column of $V_i$, i.e. $V_i^k = V_i(:,1:N_i)$. Applying the alternating iterative structure, the digital precoder $F_{BB}$ is obtained through the above steps, and the analog precoder $F_{RF}$ is updated by extracting the phase information of the $F_{opt} F_{BB}^H$ and then block diagonalized output. The steps of the Alt-ISVD algorithm is summarized in table 1.

| Algorithm 1: Alt-ISVD hybrid precoding algorithm |
|--------------------------------------------------|
| 1. Input: $F_{opt}$, $H$.                        |
| 2. Initialization: set $k = 0$, initialize the analog precoder $F_{RF}^0$, is obtained by equations (14) - (17). |
| 3. Iteration: Repeat steps 3-6 when the $k < \text{number of iterations}$. |
| 4. Fix analog precoder $F_{RF}^k$, $F_{BB}^k$ update by $F_{BB}^k = V_i U_i^H$. |
| 5. Fix digital precoder $F_{BB}^k$, take the phase angle of the $F_{opt} F_{BB}^k H$ and block diagonalized output to update $F_{BB}^{k+1}$. |
| 6. $k = k + 1$, go to step 3. |
| 7. Normalize the digital precoder, $F_{BB} = \frac{N_s}{\|F_{RF}F_{BB}\|_F} F_{BB}$. |

4. Simulation and algorithm complexity analysis

4.1. Simulation

This paper simulates a millimeter-wave MIMO system based on a PC structure in which the number of millimeter-wave channel paths is $L = 10$, and the antenna spacing of the ULA is $d = \lambda/2$, assuming that the arrival and departure azimuths angles are evenly distributed on $[0, 2\pi]$. This section presents the simulation curves of the spectral efficiency of the Alt-ISVD algorithm at low SNR and high SNR and under different RF links, and simulates the energy efficiency of the Alt-ISVD algorithm under different RF links. In the simulation, the number of transmitting antennas is $N_t = 128$, and the number of receiving antennas is $N_r = 16$ or $N_r = 32$.

In the transceiver antenna configuration is $N_t \times N_r = 128 \times 16$ and the number of transmitter RF chain is $N_{RF} = 4$, the number of data streams is $N_s = 4$, the spectrum efficiency of different algorithms under low SNR is shown in figure 2. As can be seen, the performance of Alt-ISVD algorithm is better than the Alt-SDR algorithm proposed in[11] and the algorithm 1 was proposed in[13]. As the SNR increases, the performance of the proposed algorithm is closer to that of the algorithm 2[13], but the complexity is lower than that of algorithm 1 and algorithm 2[13].
Figure 2. System performance at $N_t \times N_r = 128 \times 16$ and $N_{RF} = N_s = 4$ at low SNR.

Figure 3 shows the curves of the spectral efficiency of different algorithms increasing with the SNR at high SNR under $N_t \times N_r = 128 \times 16$ and $N_{RF} = N_s = 4$. It can be seen from the figure that the Alt-ISVD algorithm is superior to the Alt-SDR algorithm and the algorithm 1 in [13] at high SNR. When the SNR is greater than about 5dB, the spectral efficiency of the Alt-ISVD algorithm is higher than that of the algorithm 2 proposed in [13], and as the SNR increases, the performance is better than the algorithm 2 proposed in [13].

Figure 3. System performance at $N_t \times N_r = 128 \times 16$ and $N_{RF} = N_s = 4$ at high SNR.

Figure 4 shows the spectral efficiency curves of the Alt-ISVD algorithm, the FC-OMP algorithm, the full-digital precoder and the Alt-SDR algorithm for $SNR = 0$, $N_t \times N_r = 128 \times 32$ and $N_t = 2$, with different RF link numbers. It can be seen from the figure that the performance of the Alt-ISVD
algorithm is slightly lower than that of the FC-OMP algorithm. This is due to the performance loss caused by using fewer phase shifters, but the performance of the Alt-ISVD algorithm is higher than the Alt-SDR algorithm.

Figure 4. System performance in different number of RF chains for SNR = 0, \( N_t \times N_r = 128 \times 32 \) and \( N_s = 2 \)

Figure 5 shows the energy efficiency curves of the Alt-ISVD algorithm, the FC-OMP algorithm, and the Alt-SDR algorithm for SNR = 0, \( N_t \times N_r = 128 \times 32 \), and \( N_s = 2 \), under different RF link numbers. It can be seen from the figure that the energy efficiency of the Alt-ISVD algorithm is higher than the FC-OMP algorithm and the Alt-SDR algorithm. With the increase of the number of RF, the energy efficiency ratio of the Alt-ISVD algorithm and the OMP algorithm is higher. This is as the number of RF chains increases, the phase shifter that FC-OMP algorithm increases, the greatly increasing power consumption.

Figure 5. Energy efficiency in different number of RF chains for SNR = 0, \( N_t \times N_r = 128 \times 32 \) and \( N_s = 2 \)
4.2. Algorithm complexity analysis

The proposed algorithm initializes the analog precoder with the first $N_{RF}^1$ columns of the channel right singular value matrix. According to the truncated SVD (TSVD) [14], the complexity is $O \left((N_{RF}^1)^2(N_s+N_t)\right)$. The complexity of obtaining the optimal digital precoder is $O \left(N_t^2(N_s+N_t)\right)$. The complexity of updating the digital precoder and the analog precoder are $O \left(N_sN_{RF}^1N_s+N_t^2(N_s+N_{RF}^1)\right)$ and $O \left(N_sN_{RF}^1N_t\right)$ at each iteration, respectively. Algorithm 1 in[13], the initialization complexity is $O \left(N_t^1\right)$, and the complexity of each iteration is $O \left(N_t^1M^2+N_t^1M^2\right)$, where $r_t = \text{rank}(H)$, using the water-filling algorithm to calculating the digital precoder and corresponding computational complexity is $O \left(N_t^1+MN_t^1\right)$. Algorithm 2 in[13], the complexity of obtaining the analog precoder is $O \left((N_{RF}^1)^3M^2N_t^1\right)$ at each iteration, and the digital precoder is obtained by water-filling algorithm, the algorithm complexity is $O \left(N_t^1+MN_t^1\right)$. In the actual transmission system, $N_s \leq N_{RF}^1 < M < N_t < N_t^1$, it can be concluded that the complexity of the proposed Alt-ISVD algorithm is lower than the algorithms in[13].

5. Conclusions

Considering the actual engineering application, the FC hybrid precoding energy consumption is too high. In this paper, the PC hybrid precoding in millimeter-wave MIMO system is studied. It is proved that the initial analog precoder can be derived from the phase angle of the channel right singular value vector, which can avoid the complicated optimization process of searching for the optimal analog precoding matrix. It is not necessary to know the transmit antenna array response vector, which reduces the complexity, and finally alternately applies the minimum structure to improve performance. The simulation results show that with the increase of SNR, the performance of the proposed algorithm is higher than that proposed in[13], and the complexity is lower. Compared with the traditional OMP algorithm, the proposed algorithm has higher energy efficiency, lowers hardware power consumption, and is more practical.

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