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Performance Evaluation of Hybrid Precoded Millimeter Wave Wireless Communication System on Color Image Transmission

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Abstract: In this paper, a comprehensive study has been on the suitability of implementation of hybrid precoding scheme in performance analysis of the future generation wireless communication system. The 256-by-32 multi antenna supported simulated system incorporates various types of modern and classical channel coding schemes such as Low density parity check (LDPC), Repeat and Accumulate (RA), ½-rated convolutional and non-binary Bose-Chadhuri-Hocquenghem (BCH) and Zero-Forcing (ZF) signal detection technique. With consideration of ray path geometry based mmWave MIMO fading channel and properly designed precoders and combiners and their applicability in simulation works, it is seen from computer simulation results that the presently considered simulated system outperforms in retrieving color image in LDPC channel coding and 16QAM digital modulation and Zero-Forcing (ZF) signal detection schemes.

Keywords: Hybrid Precoding, Channel Coding, ZF, mmWave and Bit Error Rate (BER)

1. Introduction

The millimeter wave (mmWave) with frequency spectrum band ranging from 30GHzto300 GHz is expected to be a key component in the next generation 5G wireless communication systems. It enables the extensive use of allocated frequency spectrum to support greater data traffic for various multimedia services such as broadband mobile and backhaul services. The mmWave spectrum holds tremendous potential for providing multi-Gigabits-per-second data rates in upcoming cellular systems. One of the fundamental goals for 5G wireless mobile networks is to increase data rates through extreme densification of base stations with massive multiple-input-multiple-output (MIMO). In mmWave frequency bands, it has become a challenging task to execute cellular communication properly due to blockage, absorption, diffraction and penetration of mmWaves. However, advancement of CMOS radio-frequency technology along with the very small wavelength of mmWave signals allows for the packing of large–scale antenna arrays at both the transmit and receive ends and thus provides highly directional beam forming gains with reduced interference and acceptable signal-to-noise ratio (SNR) [1, 2]. In perspective of considering the potential of using of millimeter wave (mmWave) frequency for future 5G wireless cellular communication systems, an emphasis is being given on the study of large-scale antenna arrays for achieving highly directional beamforming. The conventional fully digital beamforming methods which require one radio frequency (RF) chain per antenna element is not viable for large-scale antenna arrays due to the high cost and high power consumption of RF chain components in high frequencies. To address the challenge of these hardware constraints, a hybrid beamforming architecture can be considered in which the overall beamformer consists of a low-dimensional digital beamformer followed by an RF beamformer implemented using analog phase shifters [3].

It is known from literature reviewing that Samsung

Electronics, an industry leader in exploring mmWave bands for mobile communications, has tested a technology that can achieve 2 Gbps data rate with 1 km range in an urban environment. Furthermore, Professor Theodore Rappaport and his research team at the Polytechnic Institute of New York University have demonstrated that mobile communications at 28 GHz in a dense urban environment such as Manhattan, NY, is feasible with a cell size of 200 m using two 25 dBi antennas, one at the BS and the other at the UE, which is readily achievable using array antennas and the beamforming technique [4].

The present study has been confined on the performance evaluative study of the simulated large-scale antenna mmWave system under consideration of hybrid beamforming structures.

2. Signal Processing Techniques

In this section, various signal processing techniques used for channel coding, geometry based fading channel estimation, hybrid precoding designing and signal detection have been outlined below.

2.1. Bose-Chadhuri-Hocquenghem (BCH) Channel Coding

BCH codes are a class of cyclic codes discovered in 1959 by Hocquenghem and independently in 1960 by Bose and Ray-Chaudhuri. The BCH codes are both binary and multi level The binary BCH code is parameterized by an integer $m \geq 3$. The t error correcting BCH code of length n is of varying nature depending on the value of m and is given by $n= 2^m-1$. Its roots include2t consecutive powers of α, the primitive element of $GF(2^m)$ [5]. In our study, a binary BCH code is of length 127 with a message is of length 64 have been used. In such [127, 64] BCH code, the value of error-correction capability, t is 10.

2.2. Low Density Parity-Check Matrix (LDPC) Channel Coding

LDPC code is a linear error correction code. Its parity check matrix H used in this paper is of 64 x 128 sized and this matrix is sparse containing less non zero elements irregularly in each row and column. The number of non zero element in each column ranges from 1 to 3and the number of non zero element in each row ranges from 5 to 6.The ½.-rated irregular LDPC code used here has a code length of 128bits.The parity-check matrix His formed from a concatenation of two matrices A and P, each with a dimension of64×64). The columns of the parity-check matrix H is rearranged to produce a modified form of parity-check matrix $\bar{H}$. With rearranged matrix elements, the matrix A becomes non-singular and it is further processed to undergo LU decomposition. The parity bits sequence p is considered to have been produced from a block based input binary data sequence $u=[u_1u_2u_3u_4\ldots\ldots u_{64}]^T$ and three matrices L,U and P(of $\bar{H}$) using the following Matlab notation:

p = mod (U\(L\z), 2);where, z = mod(P*u, 2);

The LDPC encoded 128 × 1 sized block based binary data sequencec is formulated from concatenation of parity check bit p and information bit u as: [c]=[p;u], The first 64bits of the codeword matrix [c]are the parity bits and the last 64bitsarethe information bits . The Log Domain Sum-Product LDPC is a soft decision decoding algorithm operating alternatively on the bit nodes and the check nodes through the Tanner graph. In such scheme, the received bits (0/1) are primarily converted into -1/+1 and assumed to be corrupted with AWGN channel noise of varianceσ^2 (=N0/2), N0 is the noise power spectral density. In processing, various required parameter values are compute diteratively with a view to finding out the mostly acceptable code words that satisfies the conditionc $\bar{H}^T=0$ [6, 7].

2.3. Repeat and Accumulate Channel Coding

The RA is a powerful modern error-correcting channel coding scheme. In such scheme, all the extracted binary bits from the color image has been arranged into a single block and the binary bits of the such block is repeated 2 times and rearranged into a single block containing binary data which is double of the number of input binary data[8].

2.4. Convolutional Channel Coding

In Convolutional Channel Coding, Convolutional codes are commonly specified by three parameters (n,k,m): n = number of output bits; k = number of input bits; m = number of memory registers. The quantity k/n called the code rate and it is a measure of the efficiency of the code.

The constraint length L(=k (m-1)) represents the number of bits in the encoder memory that affect the generation of the *n* output bits. Our presently considered Convolutional Channel scheme is specified with a coding rate of ½ and a constraint length of 7. The code generator polynomials G1and G2are171 and 133 in octal numbering system and can be written as [9]:

$$G1= x^0+x^2+ x^3+x^5+ x^6=1\ 0\ 1\ 1\ 0\ 1\ 1 =133 \tag{1}$$

$$G2= x^0+x^1+ x^2+x^3+ x^6=1\ 1\ 1\ 1\ 0\ 0\ 1 =171$$

2.5. Zero-Forcing (ZF) Signal Detection

In the 256 x 32 MIMO hybrid precoded system, the transmitted and received signals are represented by $X=[X_1, X_2 \ldots\ldots X_{256}]^T$ and $Y=[Y_1,Y_2 \ldots\ldots Y_{32}]^T$ respectively. If $N=[N_1, N_2 \ldots\ldots N_{32}]^T$ denotes the white Gaussian noise with a variance σ_n^2 and the channel matrix is represented by $H=[H_1H_2\ldots\ldots H_{256}]$, we can write

$$Y = \mathbf{H}X + N \tag{2}$$

As the interference signals from other transmitting antennas are minimized to detect the desired signal, the detected desired signal from the transmitting antenna with inverting channel effect by a weight matrix W is given by

$$\tilde{X} = [\tilde{X}_1, \tilde{X}_2, \ldots \tilde{X}_{256}]^T = WY \tag{3}$$

In Zero-Forcing (ZF) signal detection scheme, the ZF weight matrix is given by

$$W_{ZF} = (H^H H)^{-1} H^H \tag{4}$$

and the detected desired signal from the transmitting antenna is given by [10]

$$\tilde{X}_{ZF} = W_{ZF} Y \tag{5}$$

2.6. MIMO Fading Channel Estimation

In estimation of ray path geometry based32× 256sized mmWave MIMO fading channel H, it is assumed that the $N_t(=256)$ transmitting and $N_r(=32)$ receiving antenna sare arranged in uniform linear array (ULA).Such MIMO channel has limited scattering with Lu(=6) scatterers. Each scatterer is assumed to contribute a single propagation path between the base station (BS) and mobile station(MS). The geometrical channel model $H \in C^{N_r \times N_t}$ can be written as:

$$H = \sqrt{\frac{N_t N_r}{L_u}} \sum_{l=1}^{L_u} \alpha_{u,l} a_{MS}(\theta_{u,l}) a^*_{BS}(\varphi_{u,l}) \tag{6}$$

where, $\alpha_{u,l}$ is the complex gain of the lth path including the path loss. The variable $\theta_{u,l}$ and $\varphi_{u,l} \in [0,2\pi]$ are the lth path's angle of arrival and departure(AoAS/AoDs) respectively. Finally, $a_{BS}(\varphi_{u,l})$ and $a_{MS}(\theta_{u,l})$ are the antenna array response vectors of the BS and MS respectively.

With available knowledge of the geometry of uniform linear antenna arrays, $a_{BS}(\varphi_{u,l})$ is defined as:

$$a_{BS}(\varphi_{u,l}) = \frac{1}{\sqrt{N_t}} [1, e^{j\frac{2\pi}{\lambda} d\sin(\varphi_{u,l})}, \ldots\ldots\ldots\ldots e^{j(N_t-1)\frac{2\pi}{\lambda} d\sin(\varphi_{u,l})}]^T \tag{7}$$

And

$$a_{MS}(\theta_{u,l}) = \frac{1}{\sqrt{N_r}} [1, e^{j\frac{2\pi}{\lambda} d\sin(\theta_{u,l})}, \ldots\ldots\ldots\ldots e^{j(N_r-1)\frac{2\pi}{\lambda} d\sin(\theta_{u,l})}]^T \tag{8}$$

where, λ is the signal wavelength and d is the distance between two consecutive antenna elements.

The MIMO channel H is further normalized to get its Frobenius norm value [11, 12]

$$E[\left\| H^2_F \right\|] = N_t N_r \tag{9}$$

2.7. Precoder and Combiner Designing

The optimal unconstrained precoder F^* and optimal unconstrained combiner W^* can be estimated from implementation of singular value decomposition (SVD) to MIMO channel H in normalized form. The unconstrained RF precoder F_{RF} at the transmitter side controls phases of the up converted RF signal. Its each (i,j) th element is given by

$$F_{RF}(i, j) = \frac{1}{\sqrt{N_t}} e^{j\varphi_{i,j}} \tag{10}$$

Where $\varphi_{i,j}$ is the unquantized phase of (i, j) th element of unconstrained RF precoder F_{RF} Each entry of F_{RF} are quantized up to B bits of precision, each quantized to its nearest neighbor based on closest Euclidean distance. The phase of each entry of F_{RF} can thus be written as:

$\hat{\varphi} = (2\pi\hat{n})/(2^B)$ where, $\hat{n}$ is chosen according to

$$\hat{n} = \arg_{n \in \{0,\ldots 2^{B-1}\}} \min \left| \varphi - \frac{2\pi n}{2^B} \right| \tag{11}$$

where, φ is the unquantized phase obtained from Equation (10). Then the unconstrained RF precoder F_{RF} is computed through substituting quantized phase $\hat{\varphi}$ in Equation (11).

The optimal unconstrained precoder F^* can be written in terms of unconstrained RF precoder FRF and the uncontained baseband precoder F_{BB} as:

$$F^* = F_{RF} F_{BB} \tag{12}$$

From equation (12), we can write,

$$F_{BB} = (F_{RF}{}^T F_{RF})^{-1} F_{RF}{}^T F^* \tag{13}$$

where, $F_{RF}{}^T$ is conjugate transformed form of F_{RF}

With consideration of four RF chains, the constrained analog RF precoder F_{RF} and the constrained baseband precoder F_{BB} are estimated using the following relation based on Conjugate Gradient square method:

$$\min_{F_{RF}, F_{BB}} \left\| F^* - F_{RF} F_{BB} \right\|_F$$

$$\text{s.t.} \quad \left\| F_{RF} F_{BB} \right\|^2_F = 4 \tag{14}$$

Equation (12) can be written in modified form as:

$$F_{RF}{}^T F_{RF} F_{BB} = F_{RF}{}^T F^* \tag{15}$$

In Equation (15), the unknown F_{BB} can be determined iteratively with minimization of residual $r^{(i)} = F_{RF}{}^T F^* - F_{RF}{}^T F_{RF} F_{BB}{}^{(i)}$ using Conjugate Gradient square method [13]. The iteration terminates when the estimated residual value is $\leq 1\times10^{-10}$.

The iteratively re estimated F_{BB} value (F_{BB}) is substituted in equation (8) to get

$$F^* = F_{RF} F_{BB} \tag{16}$$

Equation (16) can be written in modified form as:

$$F_{BB}{}^T F_{RF} = F^{*T} \tag{17}$$

where, $F_{BB}{}^T$ and F^{*T} are conjugate transformed form of F_{BB}

and F* respectively.

In Equation (17), the unknown F_{RF} can be determined iteratively with minimization of residual $r^{(i)}=F^{*T}-F_{BB}^{T}F_{RF}^{(i)}$ using Conjugate Gradient square method [13]. The iteration terminates when the estimated residual value is $\leq 1\times10^{-10}$.

The iteratively re estimated F_{RF} value (F_{RF}) would be such that

$$\left\|F^{*}-\mathbf{F_{RF}}\mathbf{F_{BB}}\right\|_{F}=0 \ and \ \left\|\mathbf{F_{RF}}\mathbf{F_{BB}}\right\|_{F}^{2}=4 \quad (18)$$

In case of unconstrained RF combiner W_{RF} at the receiver side, its each (i,j) th element is given by

$$W_{RF}(i,j)=\frac{1}{\sqrt{N_r}}e^{j\varphi_{i,j}} \quad (19)$$

Each entry of W_{RF} are quantized up to B bits of precision, each quantized to its nearest neighbor based on closest Euclidean distance. The phase of each entry of W_{RF} is estimated using equation (10). The unconstrained RF combiner W_{RF} is computed through substituting quantized phase $\hat{\varphi}$in Equation (19).

The optimal unconstrained combiner W^{*} can be written in terms of unconstrained RF combiner WRF and the uncontained baseband combiner W_{BB} as:

$$W^{*}=W_{RF}W_{BB} \quad (20)$$

From equation (20), we can write,

$$W_{BB}=(W_{RF}^{T}W_{RF})^{-1}W_{RF}^{T}W^{*} \quad (21)$$

where, W_{RF}^{T} is conjugate transformed form of W_{RF} [12]

With consideration of four RF chains, the constrained analog RF combiner W_{RF} and the constrained baseband combiner W_{BB} are estimated using the following relation based on Conjugate Gradient square method:

$$\min_{W_{RF},\,W_{BB}}\left\|W^{*}-W_{RF}W_{BB}\right\|_{F}$$

$$\text{s.t.} \ \left\|W_{RF}W_{BB}\right\|_{F}^{2}=4 \quad (22)$$

Equation (20) can be written in modified form as:

$$W_{RF}^{T}W_{RF}\,W_{BB}=W_{RF}^{T}\,W^{*} \quad (23)$$

In Equation (23), the unknown W_{BB} can be determined iteratively with minimization of residual $r^{(i)}=W_{RF}^{T}W^{*}-W_{RF}^{T}W_{RF}W_{BB}^{(i)}$ using Conjugate Gradient square method [13]. The iteration terminates for the estimated residual valueis $\leq 1\times10^{-10}$.

The iteratively re estimated W_{BB}value (W_{BB}) is substituted in equation (20) to get

$$W^{*}=W_{RF}W_{BB} \quad (24)$$

Equation (24) can be written in modified form as:

$$W_{BB}^{T}W_{RF}=W^{*T} \quad (25)$$

where, W_{BB}^{T} and W^{*T} are conjugate transformed form of W_{BB} and W* respectively.

In Equation (25), the unknown W_{RF} can be determined iteratively with minimization of residual $r^{(i)}=W^{*T}-W_{BB}^{T}W_{RF}^{(i)}$ using Conjugate Gradient square method [13]. The iteration terminates when the estimated residual value is $\leq 1\times10^{-10}$.

The iteratively re estimated W_{RF} value(W_{RF}) would be such that

$$\left\|W^{*}-W_{RF}W_{BB}\right\|_{F}=0 \text{ and } \left\|W_{RF}W_{BB}\right\|_{F}^{2}=4 \quad (26)$$

In Appendix, our developed program for verifying $\left\|W^{*}-W_{RF}W_{BB}\right\|_{F}=0$ and $\left\|F^{*}-F_{RF}F_{BB}\right\|_{F}=0$ has been presented for developing idea for a typical assumed MIMO Rayleigh fading channel.

3. System Description

A simplified form of hybrid precoded millimetre wave wireless communication system is depicted in Figure 1. We consider that a color image of resolution 96 pixels (width) ×96pixels (height) will be processed in our simulated hybrid precoded millimeter wave wireless communication system.

The typically assumed color image is converted into three red, green and blue components with each component is of 96 pixels(width) ×96 pixels(height). The pixel integer values [0-255] are converted into 8 bits binary form. The binary converted signal vector $S\in(0,1)$ of dimention 1× 221184 is channel encoded and subsequently interleaved to produce a signal vector $\tilde{s}$. The transformed signal vector $\tilde{s}$ is of size1× 442368. In case of merely BCH channel coding, the signal vector $\tilde{s}$ would be of size1× 438912 and after 16-arrayQAM/PSK/DPSK digital modulation [14], the number of digitally modulated symbols is 109728 and on adding additional 864 zeros in zero padding scheme. However, 110592 × 1 sized digitally modulated signal vector $\hat{s}$ is processed in serial to paralel converter to produce blocks with each block containing N(=1024) number of digitally modulated complex symbols [$\hat{X}_0,\hat{X}_1,\hat{X}_2...........\hat{X}_{N-1}$] prior to 1024 point IDFT implementation in OFDM modulation [10]. The number of OFDM block is 108. In each OFDM block, the samples are represented by [$\hat{X}_0,\hat{X}_1,\hat{X}_2...........\hat{X}_{N-1}$] and we can write,

$$\hat{X}_k=\frac{1}{N}\sum_{n=0}^{n=N-1}\hat{X}_n e^{j2\pi nk/N}$$

$$\text{for } k=0,1,2,3,\ \ldots\ldots N-1 \quad (27)$$

The IDFT implemented 1024 × 108 sized data vector $\ddot{X}$ is reshaped into single column data vector $\ddot{X}$ of dimension 110592 × 1in parallel to serial converter and passed through Spatial demultiplexer to produce4 data stream of a4 × 27648 sized signal vector X. The signal vector X is multiplied by a digital baseband precoder matrix FBB of dimension 4× 4 and the digitally precoded signal is undergone in D/A conversion with execution of up sampling and filtering with raised cosine pulse shaping digital filter [15]. The D/A converted filtered 4 x 110632sized signal vector DA is multiplied with 256 x 4sizedanalog RF precoder FRF to produce 256 x 110632 sized signal vector FR. In RF up converter section, the signal is multiplied for each of 256 channels with multiplier ML=exp(1i*2*pi*Carrier_Freq.*t), where, Carrier_Freq is the carrier frequency in mmWave band (38GHz) and t is the sample time for each of the sample ranging from1 to 110632. A 256 x 110632 sized matrix MLL can be generated from ML using MATLAB notation MLL= repmat (ML, 256,1); The transmitted signal TX is given by TX=FR •MLL, where, •is the hadamard product which is indicative of element wise multiplication of two matrices.

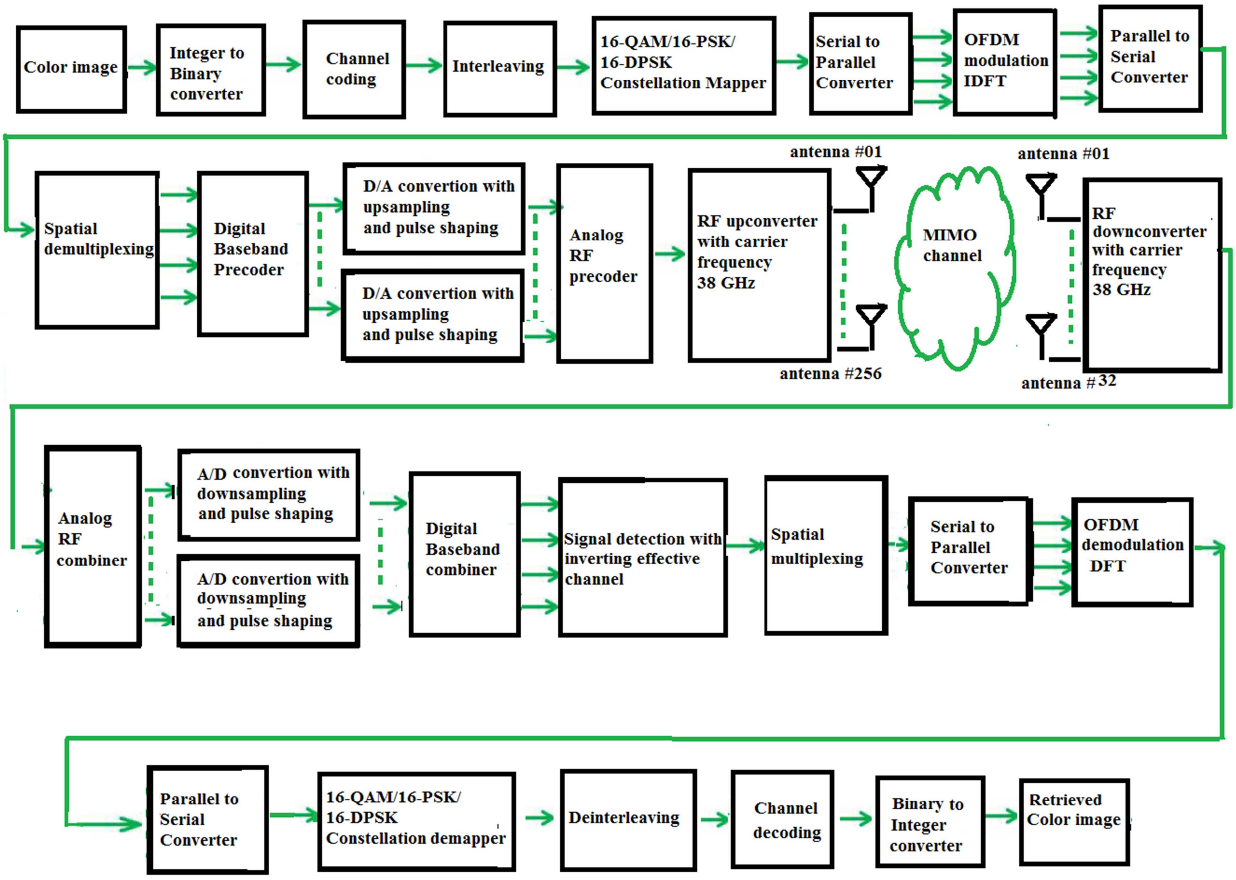

Figure 1. *Block diagram of Hybrid Precoded Millimeter Wave Wireless communication system.*

In receiving section, the 32x 110632 sized received signal Y with consideration of MIMO fading channel H is given by:

$$Y = HTX \tag{28}$$

The received signal is RF down converted through multiplication of 32x110632 sized matrix MLLL. The matrix MLLL can be generated from a matrix MLP using MATLAB notation

MLLL= repmat(MLP, 32,1) where, MLP=exp(-1i*2*pi*Carrier_Freq.*t). The received signal Y after multiplication with MLLL and in presence of addictive complex Gaussian noise N with i.i.d. CN(0, σ^2) is given by

$$\tilde{Y} = Y \bullet MLLL + N = HFRFDA + N \tag{29}$$

The noisy received signal is passed through 32 x 4 sized analog RF combiner WRF to produce modified form of received signal Y as:

$$\bar{Y} = WRF^T\tilde{Y} = WRF^T HFRFDA + WRF^T N \tag{30}$$

On executing A/D conversion theA/D converted 4 x 27648 sized received signal vector $\hat{\hat{Y}}$ is given by

$$\hat{\hat{Y}} = WRF^T HFRFFBB\bar{X} + WRF^T\bar{N} \tag{31}$$

After passing through 4 x 4 digital baseband combiner WBB, we can write,

$$\breve{Y} = WBB^T\hat{\hat{Y}} = WBB^T WRF^T HFRFFBB\bar{X} + WBB^T WRF^T\bar{N} \tag{32}$$

Where, WBB^T is the complex conjugate transformed form of WBB.

In Equation 32, it is quite observable that the effective MIMO channel $\bar{H}$ is given by

$$\bar{H} = WBB^T WRF^T HFRFFBB \tag{33}$$

On applicability of ZF based signal detection technique, the spatially demultiplexed signal $\bar{X}$ is detected. The detected signal is spatially multiplexed to produce a single column data vector $\ddot{\ddot{X}}$ of dimension 110592 × 1. In serial to parallel converter, the signal vector $\ddot{\ddot{X}}$ is reshaped into 1024 × 108 sized data vector $\bar{X}$. Ineach of 108 blocksof $\bar{X}$, the samples are represented by [$\bar{X}_0, \bar{X}_1, \bar{X}_2 .... \bar{X}_{N-1}$] and after implementation of 1024 point DFT in OFDM demodulation section, the samples in each block is represented by

$$\bar{X}_k = \sum_{n=0}^{n=N-1} \bar{X}_n e^{-j2\pi nk/N}$$

for k=0,1,2,3, ………N-1 (34)

The DFT implemented 1024× 108 sized data vector $\bar{X}$ is processed for parallel to serial conversion, digitally demodulation, de interleaving, channel decoding, binary to integer conversion to retrieve eventually the transmitted image.

4. Results and Discussion

We have conducted computer simulation study using MATLAB R2014a to evaluate the quality of the transmitted color image in a Hybrid Precoded Millimeter Wave Wireless Communication System based on the parameters presented in Table 1 It is assumed that the channel state information (CSI) of the mmWave MIMO fading channel is available at the receiver and the fading process is approximately constant during the whole period of color I mage transmission.

On critical observation of grapical illustrations presented in Figure 2 through Figure 4, it is found that the performance of the simulated systemis very much well definedunder the considered simulation parameters. In all cases, the system shows satisfactory performance in 16-QAM digital modulation and comparitively worst performancein 16-DPSKdigital modulation. In Figure 2 with consideration of ½ -rated convolutional channel coding, the ber values at 1dB SNRarefound to have values of 0.1514, 0.2182 and 0.4733 in case of 16-QAM, 16-PSK and 16-DPSK which is in dicative of system performance improvement of1.59 dBin16-QAM as compared to16-PSK and 4.95 dB in 16-QAM as compared to 16-DPSK.At 10% BER, achieved system performance improvement in terms of signal to noise ratio (SNR) are 0.5 dB and 4.8 dB in 16-QAM relative to 16-PSK and16-DPSK respectively. In Figure 3 for LDPC channel coding, the ber values at 1dB SNRare 0.0457, 0.1879 and 0.3526 for 16-QAM, 16-PSK and 16-DPSK respectively which implies system performance improvement of 6.14dB in 16-QAMrelative to 16-PSK and 8.87 dB in16-QAM relative to16-DPSK. In Figure 4 for R and A channel coding, the system performance in all considered digital modulations shows almost linear response viz. improvement of system performance occurs linearly with increase in SNR values. At a typically assumed SNR value of 1 dB, the ber valuesare 0.1705, 0.2186 and 0.2472 for 16-QAM, 16-PSK and 16-DPSK respectively which implies system performance improvement of 1.08 dB in 16-QAMrelative to 16-PSK and 1.61 dB in 16-QAMrelative to16-DPSK. In Figure 5 for BCH channel coding, the estimated ber valuesat SNR value of 1 dB are 0.1278, 0.1809 and 0.2583 for 16-QAM, 16-PSK and 16-DPSK respectively which is indicative of system performance improvement of 1.51 dB in 16-QAMrelative to16-PSKand 3.06 dB in 16-QAMrelative to 16-DPSK. In Figure 6, it is observable that the simulated system shows satisfactory performance in retrieving color image. At reasonably low SNR value, the quality of the retrieved color image is acceptable.

Table 1. *Summary of simulation model parameters.*

Parameters	Types
Data type: Color image	96 pixels(height) ×96 pixels(width)
Antenna configuration	256(Transmitting) × 32(Receiving)
Digital modulation	16-PSK and 16-QAM and 16-DPSK
Channel coding	LDPC, R and A, Convolutional and BCH
LDPC decoding Algorithm	Log Domain Sum-Product
OFDM Block Size	1024 digitally modulated symbols
OFDM symbol duration(sec)	$5.1200\times^{-05}$
Upsampling frequency(Hz)	80000000
Orthogonal subcarrier spacing (Hz)	1.9531×10^{04}
System bandwidth (MHz)	20
Sampling time for transmitted signal(sec)	1.2500×10^{-8}
Oversampling rate	4
Pulse shaping digital filter	Square root raised cosine filter
Order of filter	40
Roll of Factor of filter	0.25
Filter delay(# of input samples)	5
Carrier frequency(GHz)	28
Path loss model(dB), λ=wavelength(m)of carrier frequency, d= distance(m) between transmitter and receiver	$-20\log_{10}(\lambda/(4\pi d)$
No. of RF chains in both Transmitter and Receiver sides	4
Number of channel paths	6
Size of Baseband Precoder F_{BB}	4 × 4
Size of RF Precoder F_{RF}	256 × 4
Size Baseband Combiner W_{BB}	4 × 4
Size RF Combiner W_{RF}	32 × 4
SNR	0-10 dB
Signal detection techniques used	Zero forcing (ZF)
Channel	AWGN and Ray path based geometrical MIMO

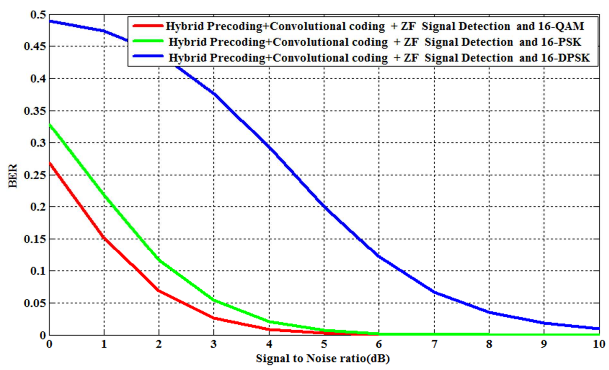

Figure 2. *BER performance comparison of Hybrid Precoded Millimeter Wave Wireless Communication system with implementation of ZF signal detection, ½ -rated convolutional channel coding and various digital modulation schemes.*

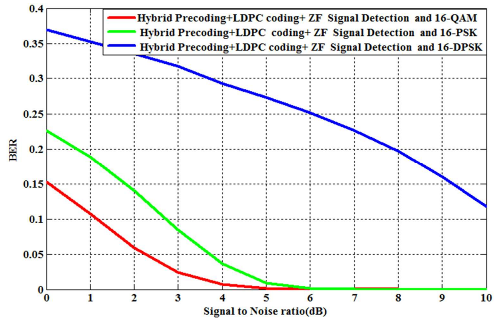

Figure 3. *BER performance comparison of HybridPrecoded Millimeter Wave Wireless Communication system with implementation of ZF signal detection, LDPC channel coding and various digital modulation schemes.*

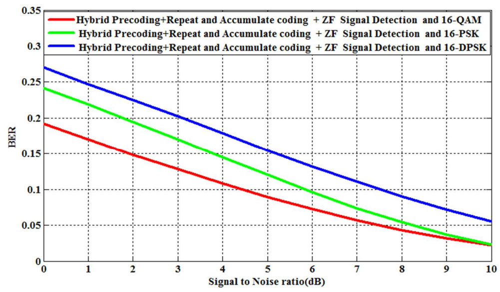

Figure 4. *BER performance comparison of Hybrid Precoded Millimeter Wave Wireless Communication system with implementation of ZF signal detection, Repeat and Accumulate channel coding and various digital modulation schemes.*

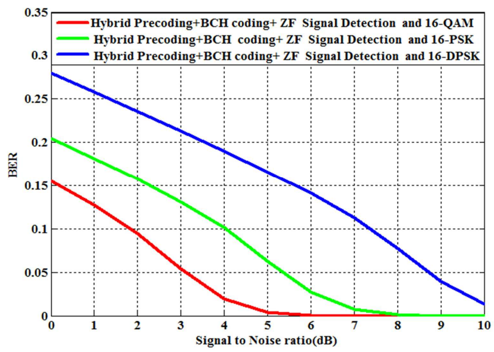

Figure 5. *BER performance comparison of Hybrid Precoded Millimeter Wave Wireless Communication system with implementation of ZF signal detection, BCH channel coding and various digital modulation schemes.*

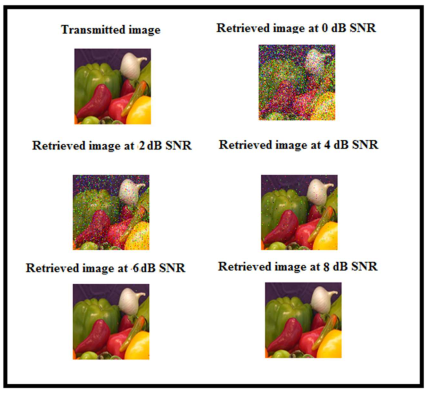

Figure 6. *Transmitted and retrieved images in Hybrid Precoded Millimeter Wave Wireless Communication system under implementation of ZF signal detection, LDPC channel coding and 16-QAM digital modulation scheme.*

5. Conclusions

In this present paper, we have made a comprehensive study on the performance analysis of millimeter wave (mmWave) wireless communication system under simultaneous implementation of both digital and analog precoding and combining schemes in hybrid form. Simulation results ratify that the Hybrid precoders and combiners have been designed satisfactorily with the application of effective iterative Conjugate Gradient Squared (CGS) Method and Singular value decomposition of ray path geometry based mmWave MIMO fading channel. Based on the results presented in this paper on color image transmission, it can be concluded that the presently considered Hybrid Precoded Millimeter Wave wireless communication System is undoubtedly a robust system in perspective of signal transmission in hostile fading channel under implementation of LDPC channel coding, 16 QAM digital modulation and Zero Forcing (ZF) signal detection schemes.

Appendix

```
clear all; close all;
%antenna configuration: 32 receiving × 256 transmitting
%MIMO fading channel generation
H=sqrt(1/2)*(randn(32,256)+sqrt(-1)*randn(32,256));
% MIMO channel
% Normalization of channel matrix
for kk=1:256
for kkk=1:32
H(kkk,kk)=H(kkk,kk)/ (abs(H(kkk,kk)));
end; end;
channel_normalization=(norm( H,'fro')).^2 ;
% its value would be 32 x 256=8192
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
% Singular value decomposition (SVD) of channel matrix H
[U SIGMA VT] = svd(H); % U: 32 x 32
V=VT'; % 256 x 256
FSTAR=V(:,1:4); % 256 x 4, optimal unconstrained precoder
WSTAR=U(:,1:4); % 256 x 4 optimal unconstrained
combiner
%FSTAR=Baseband Precoder(FBB) X RF Precoder(FRF)
%WSTAR=Baseband Combiner(WBB) X RF
Combiner(WRF)
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%%
% Unconstrained FRF and FBB estimation
% no of stream =4, FRF= 256 x 4
quantized_phase=((0:2^7-1)*2*pi/2^7)';
unquantized_phase=2*pi*rand(256,4); % 256 rows x4 cols
for kk=1:4
for kkk=1:256
for kkkk=1:128
nhat(kkk,kk,kkkk)= abs((unquantized_phase(kkk,kk)-
quantized_phase(kkkk)));
end; end;end;
for kk=1:4
```

```
for kkk=1:256
[value(kkk,kk), integer(kkk,kk)]= min(nhat(kkk,kk,:));
phihat(kkk,kk)=2*pi*integer(kkk,kk)/(2^7);
end; end;
%%%%%%%%%%%%%%%%%%%%%%%
FRF= (1/sqrt(256))*exp(j*phihat); % 256 x 4 RF Precoder
%%%%%%%%%%%%%%%%%%%%%%%%%%%
FBB=inv(FRF'*FRF)*FRF'*FSTAR;
%%%%%%%%%%%%%%%%%%%%%%%%%%%
% Reestimate FRF and FBB such that FRF*FBB=FSTAR
modified_FRF=FRF'*FRF;
modified_FSTAR=FRF'*FSTAR;
%Conjugate Gradient  square Method
for kk=1:4
Estimated_FBB(:,kk)=
cgs(modified_FRF,modified_FSTAR(:,kk), 1e-10);
end;
%FBB'*FRF' =FSTAR', FSTAR'= 8 rows x 256 cols FRF' =
8 rows x 256 cols
FSTART=FSTAR';
FBBT=Estimated_FBB';
for kk=1:256
Estimated_FRF(:,kk)= cgs(FBBT,FSTART(:,kk), 1e-10);
end;
Estimated_FRF=Estimated_FRF';
% Restimate WRF and WBB such that WRF*WBB=WSTAR
unquantized_phase1=2*pi*rand(32,4); % 32 rows x 8 cols
% From Low-Complexity Hybrid Precoding equation 6,
quantized phase
for kk=1:4
for kkk=1:32
for kkkk=1:128
nhat1(kkk,kk,kkkk)= abs((unquantized_phase1(kkk,kk)-
quantized_phase(kkkk)));
end;end;end;
%%%%%%%%%%%%%%%%%%%%%%%%%%
for kk=1:4
for kkk=1:32
[value1(kkk,kk), integer1(kkk,kk)]= min(nhat1(kkk,kk,:));
phihat1(kkk,kk)=2*pi*integer1(kkk,kk)/(2^7);
end;end;
%%%%%%%%%%%%%%%%%%%%%%%
WRF= (1/sqrt(32))*exp(j*phihat1); % 32 x 4 RF Precoder
WBB=inv(WRF'*WRF)*WRF'*WSTAR;
%%%%%%%%%%%%%%%%%%%%%%%%%%%%
%
modified_WRF=WRF'*WRF;
modified_WSTAR=WRF'*WSTAR;
% Equation modified_WRF*WBB=modified_WSTAR
%Conjugate Gradient square Method
for kk=1:4
Estimated_WBB(:,kk)=
cgs(modified_WRF,modified_WSTAR(:,kk), 1e-10);
end;
%WBB'*WRF' =WSTAR', WSTAR'= 8 rows x 256 cols
WRF' = 8 rows x 256 cols
WSTART=WSTAR';
WBBT=Estimated_WBB';
for kk=1:32
Estimated_WRF(:,kk)= cgs(WBBT,WSTART(:,kk), 1e-10);
end;
Estimated_WRF=Estimated_WRF';
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
result1= round(norm(FSTAR-
Estimated_FRF*Estimated_FBB,'fro'))
result2= round(norm(WSTAR-
Estimated_WRF*Estimated_WBB,'fro'))
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
```

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