

Efficient Limited Feedback Technique for FDD MIMO Systems

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ARTICLE INFO

Article history:

Received: 15 November, 2017

Accepted: 05 February, 2018

Online: 17 March, 2018

Keywords:

MIMO OFDM

Beamforming

Feedback overhead

ABSTRACT

In this paper an efficient feedback quantization technic for beamforming in MIMO systems is presented. The proposed technic named time domain quantization TD-Q is based on the feedback of time domain parameters necessary for the reproduction of the beamforming matrix at the transmitter. This TD-Q presents the same performance than the conventional Givens rotation quantization GR-Q approach which is adopted in IEEE 802.11ac standard. The performance and amount of feedback of the proposed TD-Q are studied and compared with the GR-Q in IEEE 802.11ac context.

1 Introduction

The combination of Multiple-Input Multiple-Output (MIMO) and Orthogonal Frequency Division Multiplexing (OFDM) technologies (MIMO-OFDM) is now adopted in several communication standards, including the 5th generation of mobile communication network (5G) [1], the IEEE WLAN 802.11ac [2] and the IEEE 802.16 standards (WiMax)[3].

On the one hand, OFDM is a worthwhile trade-off between bit-error rate performance and spectral efficiency. OFDM consumes part of the channel bandwidth, but it is robust to frequency selective fading environment. In addition, it enables the use of several advanced technics to further enhance the system throughput, as for instance the bit loading technic [4] and the subcarriers allocation in orthogonal frequency division multiple access (OFDMA) [5].

On the other hand, the MIMO system has the potential to improve the system capacity. In IEEE 802.11ac wireless local area network (WLAN) standard, there are five transmit and receive MIMO technics: Cyclic Shift Diversity (CSD), Space Time Block Coding (STBC), Spatial Division Multiplexing (SDM), Maximal Rotation Combining (MRC) and Transmit Beamforming (TxBF) [2]. Among these five technics, this is the Transmit Beamforming which can maximize the system capacity. Indeed, the Beamforming with precoding and postcoding eliminates the co-channel interferences (CCI) which are the fundamental problem faced by the practical MIMO system [6-8].

However, in beamforming technic the channel state information (CSI) must be available at the transmitter

and the receiver. This CSI is estimated at the receiver and fed back to the transmitter. The CSI, which indicates amplitude and phase for each transmit antenna, receive antenna, and each OFDM subcarrier in the RF channel may reduce the overall throughput. To reduce the feedback amount, the receiver computes the beamforming matrix (precoding matrix) which is a unitary matrix and compresses it before sending back to the transmitter.

There are several proposals in the literature for the beamforming matrix compression. Many authors propose codebook based approach. In codebooks based technics, the channel distribution is taken into account during the codebook design. Instead of sending the precoding matrix, only the index of the selected precoding matrix (after channel estimation and SVD decomposition) is sent. Intrinsically, the codebook is restricted to have fixed cardinality. Thereby, the selected pre-coding matrix, which is the most similar is not necessarily the most optimal. Unfortunately, the codebook size has an impact on the system performance and requires high storage. For traditional MIMO systems, several codebooks have been proposed, such as Kerdock codebook [9], codebooks based on vector quantization [10], Grassmannian packing [11], discrete Fourier transform (DFT) [12] and quadrature amplitude modulation [13]. The codebook principle is adopted in LTE [14]. In [15], the codebook was designed and optimized by selecting matrices having 8 phase-shift keying (PSK) entries and coping with a large range of propagation conditions. However the cost of this scheme is that the performance gain decreases dramatically due to the channel quantized er-

ror caused by the codebook feedback[16].

Another alternative is the Givens rotation (GR) approach. Here, Givens rotation is used to decompose the unitary beamforming matrix. After the decomposition, the receiver only feeds back the GR parameters necessary for the reproduction of the beamforming matrix by the transmitter [17]. The GR approach is adopted in IEEE 802.11ac standard for TxBF mode[2]. The compression ratio of the GR quantization is 75% for a MIMO 2×2 configuration [18]. However, this ratio decreases depending on the number of antennas. It then becomes necessary to find an alternative to the GR quantization since the 802.11ac standard has adopted an 8×8 MIMO configuration which may reach 10Gbps and this number of antennas will undoubtedly increase due to the growing demand for bitrate.

This paper proposes a quantization approach named time domain quantization (TD-Q). Instead of the quantized precoder matrix as in GR, we propose the feedback of the quantized time domain channel coefficients. In addition, an optimization based on a metric updated during the channel estimation process and the MIMO channels impulse response is also proposed in order to further reduce the feedback overhead.

The rest of the paper is organized as follows. The section 2 presents the beamforming in MIMO OFDM systems. Section 3 is dedicated to the GR-Q which is adopted in the IEEE 802.11ac standard. Following that, the proposed TD-Q is presented in the section 4. Next, simulation results and comparison between GR-Q and TD-Q are shown in the section 5. Finally, the conclusion on the works presented in this paper is done in the section 6.

2 System Model

Consider a MIMO channel with N_t transmit antennas and N_r receiver antennas. In classical MIMO-OFDM context, for each subcarrier, the $N_r \times 1$ received signal y can be expressed as:

$$y = Hx + n \quad (1)$$

where H is the $N_r \times N_t$ frequency channel coefficients, x the $N_t \times 1$ transmitted signal and n the $N_r \times 1$ zero mean Gaussian noise vector.

By using singular value decomposition (SVD), the matrix H can be decomposed into:

$$H = UDV^H \quad (2)$$

where $U \in \mathbb{C}^{N_r \times R}$ and $V \in \mathbb{C}^{N_t \times R}$ are unitary matrix, $D \in \mathbb{C}^{R \times R}$ is diagonal, $(.)^H$ denotes the hermitian of $(.)$ and R the rank of the matrix H .

In MIMO-OFDM with beamforming context as in TxBF mode of IEEE 802.11ac, the received signal y is expressed this time by:

$$y = U^H H V x + U^H n = U^H U D V^H V x + U^H n = D x + U^H n \quad (3)$$

where V and U^H are used as a precoder matrix and a postcoder matrix respectively. The postcoder matrix

U^H is an unitary matrix and thus the noise is neither colored nor enhanced.

Since D is exactly diagonal, there is no CCI and the MIMO channel matrix is reduced on R separate and independent SISO channels.

However, the precoder matrix V is needed at the transmitter. Two solutions are available in order to provide CSI at the transmitter. The first solution consists of exploiting the channel reciprocity in case of time division duplex (TDD) transmission modes. But, it is not straightforward due to the mismatch in the radio front end. The alternative, which is the most appropriate and that can be applied for both TDD and frequency division duplex (FDD) transmission modes, consists in using the reverse link to feed back the matrix V to the transmitter. Note that, for each subcarrier the unitary matrix V has to be quantized and fed back to the transmitter.

3 Givens Rotation Quantization and feedback compression ratio analysis

3.1 Givens Rotation Quantization principle

Givens Rotation Quantization GR-Q is proposed in [17]. The authors propose a reduction of the feedback overhead by exploiting the unitary property of the precoder matrix V . The idea is to represent the matrix V as a special form of Givens rotation with a complex diagonal matrix. Thereby, the $N_t \times R$ unitary matrix V can be factorized as follows.

$$V = \prod_{i=1}^{\min(R, N_t-1)} [D_i \prod_{l=i+1}^{N_t} G_{li}^T(\psi_{li})] \times I_{N_t \times R} \quad (4)$$

where

$$D_i = \begin{bmatrix} I_{i-1} & 0 & 0 & \dots & 0 \\ 0 & e^{j\phi_{i,i}} & 0 & \dots & 0 \\ 0 & 0 & \dots & 0 & 0 \\ 0 & 0 & 0 & e^{j\phi_{N_t-1,i}} & 0 \\ 0 & 0 & 0 & 0 & 1 \end{bmatrix} \quad (5)$$

$$G_{li} = \begin{bmatrix} I_{i-1} & 0 & 0 & \dots & 0 \\ 0 & \cos(\psi_{li}) & 0 & \sin(\psi_{li}) & 0 \\ 0 & 0 & I_{l-i-1} & 0 & 0 \\ 0 & -\sin(\psi_{li}) & 0 & \cos(\psi_{li}) & 0 \\ 0 & 0 & 0 & 0 & I_{N_t-l} \end{bmatrix} \quad (6)$$

The matrix D_i is an $N_t \times N_t$ diagonal matrix, $I_{N_t \times R}$ is an $N_t \times R$ identity matrix and G_{li} is an $N_t \times N_t$ Givens rotation matrix.

According to the equations (4-6), the parameters to be determined to identify the precoder matrix V are:

$$\begin{aligned} \psi_{li} & \text{ for } i = 1, 2, \dots, m \text{ and } i < l \leq N_t \\ \phi_{j,i} & \text{ for } i = 1, 2, \dots, m \text{ and } i < j \leq N_t - 1 \end{aligned} \quad (7)$$

where $m = \min(R, N_t - 1)$.

Take a 3×3 unitary matrix V as an example, it can be expressed as:

$$V = D_1 \times G_{21}^T(\psi_{21}) \times G_{31}^T(\psi_{31}) \times D_2 \times G_{32}^T(\psi_{32}) \times I_{3 \times 3} \quad (8)$$

The precoder matrix V can then be reconstructed through the six parameters $\psi_{21}, \psi_{31}, \psi_{32}, \phi_{1,1}, \phi_{2,1}$ and $\phi_{2,2}$.

Therefore, instead of all the elements of the matrix V , it is sufficient to consider the parameters ψ_{li} and $\phi_{j,i}$. These parameters can vary from 0 to 2π for ϕ and from 0 to $\pi/2$ for ψ .

Now, ψ and ϕ can be quantized according to equations (9) and (10) where $\widehat{\psi}$ and $\widehat{\phi}$ represent the quantized angles.

$$\widehat{\psi} = \frac{k\pi}{2^{nb_\psi+1}} + \frac{\pi}{2^{nb_\psi+2}} \quad (9)$$

where $k = 1, 2, \dots, 2^{nb_\psi} - 1$ and nb_ψ the number of bits per angle ψ

$$\widehat{\phi} = \frac{k\pi}{2^{nb_\phi-1}} + \frac{\pi}{2^{nb_\phi}} \quad (10)$$

where $k = 1, 2, \dots, 2^{nb_\phi} - 1$ and nb_ϕ the number of bits per angle ϕ

Finally, the receiver feeds back the quantized parameters $\widehat{\psi}_{li}$ and $\widehat{\phi}_{j,i}$ to the transmitter which can recover the quantized precoder matrix \widehat{V} by using:

$$\widehat{V} = \prod_{i=1}^{\min(R, N_t-1)} [\widehat{D}_i \prod_{l=i+1}^{N_t} \widehat{G}_{li}^T(\widehat{\psi}_{li})] \times I_{N_t \times R} \quad (11)$$

where

$$\widehat{D}_i = \begin{bmatrix} I_{i-1} & 0 & 0 & \dots & 0 \\ 0 & e^{j\widehat{\phi}_{i,i}} & 0 & \dots & 0 \\ 0 & 0 & \dots & 0 & 0 \\ 0 & 0 & 0 & e^{j\widehat{\phi}_{N_t-1,i}} & 0 \\ 0 & 0 & 0 & 0 & 1 \end{bmatrix} \quad (12)$$

It is important to note that:

- The receiver has to feed back a precoder matrix \widehat{V} for each subcarrier.
- The quantization error Σ with $\widehat{V} = V + \Sigma$ exists on each subcarrier.

Consider the equation (3) with the quantized error Σ and \widehat{V} as a precoder matrix.

$$\begin{aligned} y &= U^H U D V^H \widehat{V} x + U^H n \\ &= U^H U D V^H (V + \Sigma) x + U^H n \\ &= D x + \widehat{D} V^H \Sigma x + U^H n \\ &= D x + M_{CCI} x + U^H n \end{aligned} \quad (13)$$

The term $M_{CCI} x = (D V^H \widehat{V} - D) x$ is the corresponding CCI caused by the quantization error. However, in telecommunication standard, one always chooses a number of quantization bits (nb_ψ and nb_ϕ) that allows

the minimization of resulting CCI. This is the case of IEEE 802.11ac where $nb_\psi = nb_\phi = 10$.

Finally, the feedback overhead length (in bit) noted by N_{FGR} can be calculated using the formula $N_{FGR} = N_u * nb * N_t (N_t - 1)$ assuming that $N_r = N_t$, $nb_\psi = nb_\phi = nb$ and N_u the number of used subcarriers.

3.2 Feedback overhead length analysis in IEEE 802.11ac context

The bandwidth in IEEE802.11ac has increased from a maximum of 40 MHz with the old standard up to 80 or even 160 MHz. The number of subcarriers and used one according to the bandwidth are grouped in the Table 1.

Table 1: Number of used subcarriers according to the de bandwidth in IEEE 802.11ac

Bandwidths	N	N_u
20 MHz	64	52
40 MHz	128	108
80 MHz	256	234
160 MHz	512	486

Table 2 contains for any MIMO configurations and bandwidth:

- the feedback overhead length when no compression is performed,
- the feedback overhead length when GR quantization is performed,
- the compression ratio achieved by GR quantization.

As we can see, the GR quantization can reach a compression ratio up to 75% for a MIMO 2×2 configuration. This compression ratio decreases when the number of antennas becomes large. Because of the high demand in terms of transmission throughput, the number of antennas will increase in the evolution of IEEE 802.11 standard. The next WLAN generation will undoubtedly consider the massive MIMO concept which is more and more studied for WLAN [19-20]. In this context, it is necessary to find an alternative to the GR quantization because its compression ratio, as demonstrated by the equation (14), will be around 50%:

$$\lim_{N_t=N_r \rightarrow +\infty} \frac{N_t(N_t-1)}{N_t * N_t * 2} = \frac{1}{2} \quad (14)$$

4 Time domain quantization

4.1 Time domain quantization principle

Consider a MIMO channel with N_t transmit antennas and N_r receive antennas. The discrete time domain channel response between the transmitting antenna i

Table 2: Feedback overhead (in bit) with and without GR quantization for any configurations ($N_t \times N_r$ and bandwidth).

MIMO configuration	Number of GR parameters	feedback overhead N_{FGR} for GR				feedback overhead without quantization				Ratio
		20 MHz	40 MHz	80 MHz	160 MHz	20 MHz	40 MHz	80 MHz	160 MHz	
2x2	2	1040	2160	4680	9720	4160	8640	18720	38880	75%
3x3	6	3120	6480	14042	29160	9360	19365	42120	87480	66.67%
4x4	12	6240	12960	28080	58320	16640	34560	74880	155520	62.5%
6x6	30	15600	32400	70200	145800	37440	77760	169730	349920	58.33%
8x8	56	29120	60480	131040	271410	53248	110592	239616	622080	56.25%

and the receiving antenna j under the multipath fading environments can be expressed as:

$$h_{ij}(n) = \sum_{l=0}^{L-1} h_{ij,l} \delta(n - \tau_{ij,l}) \quad (15)$$

where L is the number of paths, $h_{ij,l}$ and $\tau_{ij,l}$ the complex time varying channel coefficient and delay of the l^{th} path.

Note that in practical OFDM system, in order to eliminate the intersymbol interference (ISI) between consecutive OFDM symbols, the maximum multipath delay is within the cyclic prefix, of length (CP) ie $L \leq CP$. Thereby, we propose, instead of the quantized precoder matrix V for each subcarrier, to feedback the quantized time domain channel coefficients for the entire system. The main goal of this proposition is to considerably reduce the feedback overhead because in this case, the feedback is composed at most by $N_t * N_r * CP$ complex coefficients. As can be seen, the number of OFDM subcarriers has no impact on the feedback overhead length. In addition, this proposed time domain quatization method can be performed jointly with the channel estimate.

4.2 Joint time domaine channel estimation and feedback quantization

In closed loop MIMO-OFDM systems, a preamble is inserted at the beginning of frames to facilitate the channel estimation. Comb-type pilot method can also be used on the rest of the frame (as in IEEE 802.11ac [2] or LTE [1]) to improve the accuracy of the channel estimation during the transmission. In this paper, we focus on time domain channel estimation (TD-CE). This technique is known to provide very good results by significantly reducing the noise on the estimated channel coefficients [21][22].

Firstly, without using any knowledge of the statistics on the channels, least square (LS) channel estimation can be performed at the receiver side using the demodulated pilots signal and the known pilots symbol in the frequency domain for each subcarrier.

Next assuming that $H_{ij,k}$ is the discrete LS estimated channel response on subcarriers k between the i^{th} transmit antenna and the j^{th} receive antenna, the CSI can be converted into the time domain by the in-

verse discrete fourier transform (IDFT) algorithm as:

$$h_{ij}(n) = IDFT(H_{ij}) = \sqrt{\frac{1}{N_s}} \sum_{k=0}^{N_s-1} H_{ij,k} e^{j\frac{2\pi n k}{N_s}} \quad (16)$$

where N_s is the number of subcarriers and $n = 1, \dots, N_s$.

Note that the LS method is widely used due to its simplicity and minimum requirements for the knowledge of channel statistics. However, the LS estimator does not consider the noise effect. That is why its performance is often degraded by the noise. By taking into account the equation (15) and the fact that $L \leq CP$, the time domain samples $h_{ij}(n)$ from ((16)) can be divided into two parts:

- the first CP samples in which are the paths of the channel,
- the other samples which are only composed by noise.

A smothing process which keeps only the first CP samples by eliminating the other one can considerably improve the accurate of the estimated channel [21]. However, it is possible to estimate a mean of the noise noted by \bar{m} from the noise samples before its deletion.

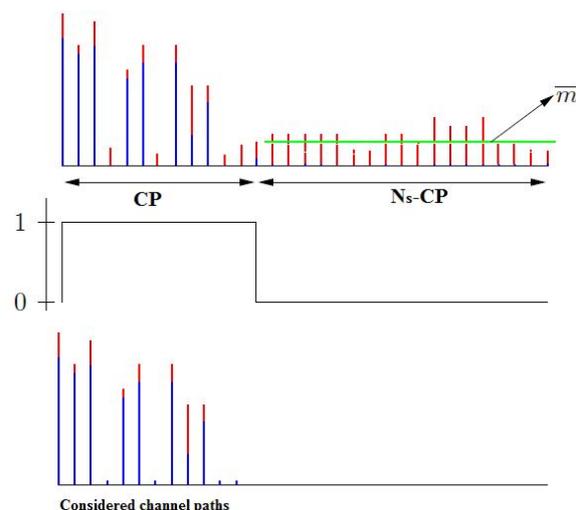


Figure 1: Mean Noise

This proposed process is illustrated in the figure 1. the metric \bar{m} estimated from the noise samples allows to keep only the channel paths which are among the

first CP samples. As can be seen, all samples below the \bar{m} were eliminated.

Now, the receiver have to do the two following actions:

- Perform a DFT of size N_s by considering only the remaining samples $h_{ij}(n)$. The resulting estimated channel frequency response is then more accurate than the LS one.
- Perform the feedback of the remaining samples $h_{ij}(n)$. To achieve this goal, each of the remaining samples $h_{ij}(n)$ whose number is lower than CP is expressed as:

$$h_{ij}(n) = \rho_{ij} e^{j\phi_{ij,n}} = \tan(\psi_{ij,n}) e^{j\phi_{ij,n}} \quad (17)$$

where $n \leq CP$, $\psi_{ij,n}$ and $\phi_{ij,n}$ angles parameters can vary from 0 to $\pi/2$ and 0 to 2π respectively.

After that, $\psi_{ij,n}$ and $\phi_{ij,n}$ is quantized and fed back to the transmitter according to equations (9) and (10) where $\widehat{\psi}_{ij,n}$ and $\widehat{\phi}_{ij,n}$ represent the quantized angles parameters.

Thus, the feedback overhead length (in bit) noted by N_{FTD} can be calculated using the formula $N_{FTD} = N_{rs} * 2 * nb * N_t^2$ assuming that $N_r = N_t$, $nb_\psi = nb_\phi = nb$ and N_{rs} the number of remaining samples.

Thanks to this feedback, the transmitter can reconstitute the accurate estimated channel. Thereby, the MIMO channel response is available both at the reciever and the transmitter side. Finally, transmitter and reciever perform SVD decomposition which allows it to respectively have the pre-coder and the post-coder.

5 Performance comparison in IEEE 802.11ac context

In this section, we compare the Givens rotation quantization GR-Q and the proposed time domain quantization TD-Q one in terms of feedback overhead and BER performance in IEEE 802.11ac context [2]. Note that the bandwidth in IEEE802.11ac has increased from a maximum of 40 MHz with the old standard up to 80 or even 160 MHz. For each bandwidth, the cyclic prefix length (CP) is equal to a quarter of the number of subcarriers (in Table 1). We consider in the simulations MIMO 2×2 , 3×3 , 4×4 , 6×6 and 8×8 with 4-QAM and 64-QAM modulations.

In the sequel, all simulations are performed using the ITU Channel Model for Outdoor to Indoor and Pedestrian Test Environment. Since the delay spread can vary significantly, the ITU recommendation specifies two different delay spreads for each test environment: low delay spread (channel model A), and medium delay spread (channel model B). The parameters of these channel's model are grouped in Table 3. In this paper we focus on the channel model B.

5.1 feedback overhead comparison

Figures 2 to 6 show the evolution of the total number of feedback complex coefficients of the proposed time domain quantization according to the signal to noise ratio (SNR) for MIMO 2×2 , 3×3 , 4×4 , 6×6 and 8×8 configurations. For each MIMO configuration, simulations are performed for the 4 existing bandwidth of the IEEE 802.11 ac standard (20 MHz,40 MHz,80 MHz and 160 MHz). Note that the total number of feedback complex coefficients represents the sum of all the remaining samples and it can be calculated by $N_{rs} * N_t^2$ assuming that $N_r = N_t$.

As can be seen from these figures:

- whatever the MIMO configuration, the total number of feedback complex coefficients increases according to the bandwidth and the number of antennas.
- whatever the SNR, the plotted cumulative distribution function (CDF) shows that it is statistically possible for each MIMO configuration and bandwidth to know the maximum value of the total number of feedback complex coefficients. Following this study the total number of feedback complex coefficients for any configurations ($N_t \times N_r$ and bandwidth) in IEEE 802.11ac ontext are grouped in the table 4.

Therefore, according to equation (17) the feedback overhead length of the proposed time domaine quati-zation (in bit) noted by N_{FTD} can be calculated using $N_{FTD} = N_{rs} * 2 * nb * N_t^2$ and the number of remaining samples N_{rs} in the Table 4.

Except for MIMO 2×2 configuration and bandwidth 20MHz, the feedback overhead of the proposed time domain quantization is much shorter than that of the Givens Rotation quantization. The compression ratio of the proposed time domain quantization compared to Givens Rotation quantization can be up to almost 70%.

The most important to note here is the fact that unlike the Givens Rotation quantization for which the compression ratio decreases when the number of antennas becomes large (see section 3.2), the proposed time domain compression becomes more efficient. This last remark makes the proposed time domain compression very interesting for the next WLAN generation which will undoubtedly consider the massive MIMO concept which is more and more studied for WLAN [19-20].

Table 3: ITU Channel Model for Outdoor to Indoor and Pedestrian Test Environment.

Tap	Channel A		Channel B	
	Relative delay(ns)	Average power (dB)	Relative delay(ns)	Average power (dB)
1	0	0	0	0
2	110	-9.7	220	-0.9
3	190	-19.2	800	-4.9
4	410	-22.8	1200	-8
5			2300	-7.8
6			3700	-23.9

Table 4: Total number of feedback complex coefficients of the proposed time domain quantization for any configurations ($N_t \times N_r$ and bandwidth) in IEEE 802.11ac context.

MIMO configuration	feedback overhead N_{FTR}			
	20 MHz	40 MHz	80 MHz	160 MHz
2x2	55	90	161	293
3x3	111	192	334	613
4x4	191	323	582	1069
6x6	415	703	1254	2337
8x8	718	1215	2205	4106

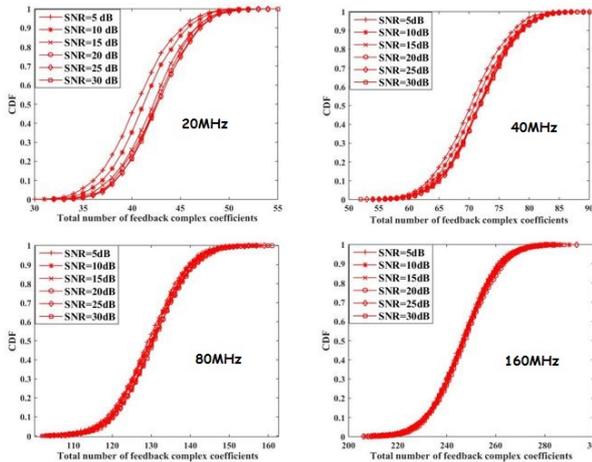


Figure 2: The Cumulative Distribution Function of the total number of feedback complex coefficients according to the Signal to Noise ratio and Bandwidth for MIMO 2×2 configuration

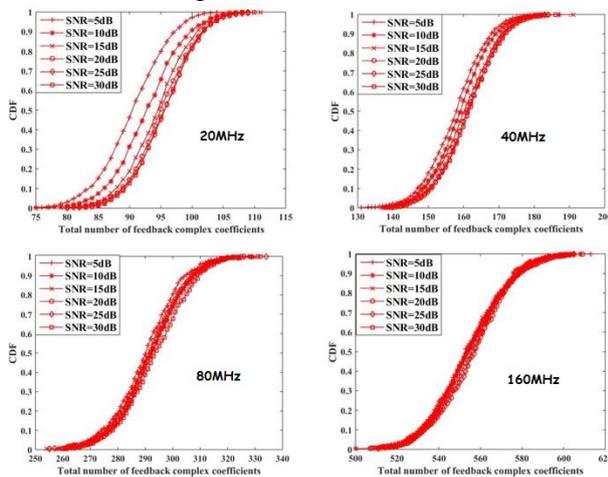


Figure 3: The Cumulative Distribution Function of the total number of feedback complex coefficients according to the Signal to Noise ratio and Bandwidth for MIMO 3×3 configuration

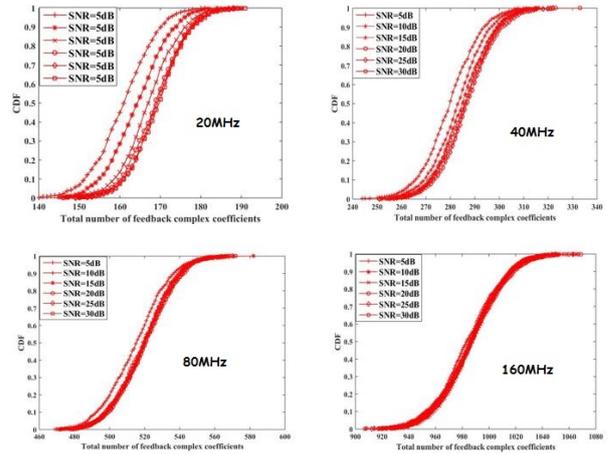


Figure 4: The Cumulative Distribution Function of the total number of feedback complex coefficients according to the Signal to Noise ratio and Bandwidth for MIMO 4×4 configuration

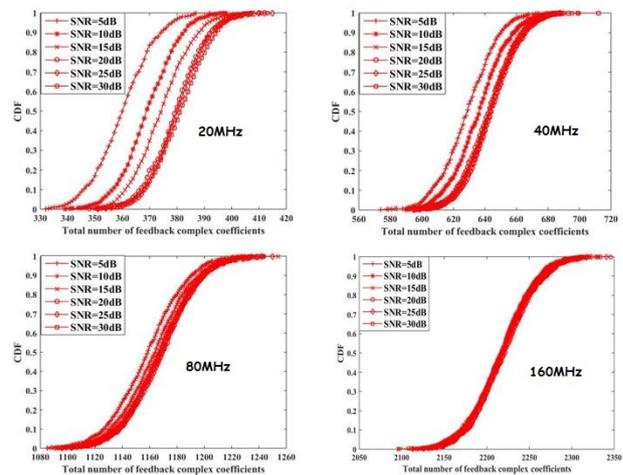


Figure 5: The Cumulative Distribution Function of the total number of feedback complex coefficients according to the Signal to Noise ratio and Bandwidth for MIMO 6×6 configuration

5.2 BER performance

In this subsection, we compare the beamforming with Givens rotation quantization (GR-Q) and that with the time domain quantization in terms of bit error rate (BER) performance. The considered simulation parameters are listed in Table 6.

Table 5: Feedback overhead (in bit) comparison between Givens Rotation quantization and the proposed time domain quatization for any configurations ($N_t \times N_r$ and bandwidth).

MIMO configuration	feedback overhead N_{FGR} for Givens Rotation quantization				feedback overhead N_{FTR} for the proposed time domain quatization				Ratio	
	20 MHz	40 MHz	80 MHz	160 MHz	20 MHz	40 MHz	80 MHz	160 MHz	min	max
2x2	1040	2160	4680	9720	1100	1800	3220	5860	-0.5%	39.7%
3x3	3120	6480	14042	29160	2220	3840	6680	12260	28.8%	58%
4x4	6240	12960	28080	58320	3820	6460	11640	21380	38.7%	63.3%
6x6	15600	32400	70200	145800	8300	14060	25080	46740	46.8%	68%
8x8	29120	60480	131040	271410	14360	24300	44100	82120	50.6%	69.7%

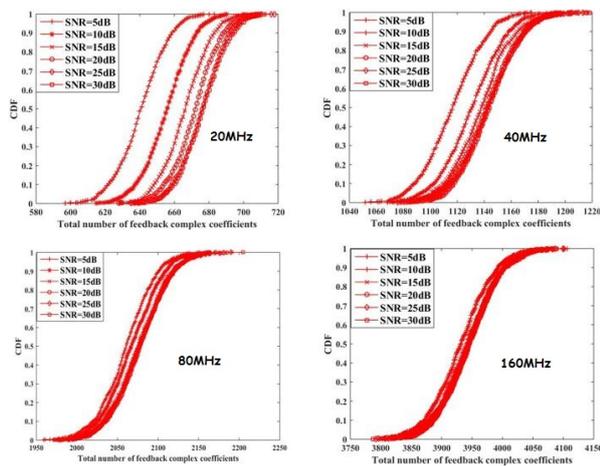


Figure 6: The Cumulative Distribution Function of the total number of feedback complex coefficients according to the Signal to Noise ratio and Bandwidth for MIMO 8×8 configuration

Table 6: Simulation Parameters

Channel Model	ITU Channel Model
Bandwidth	40MHz
Number of FFT points (N)	128
Number used subcarriers	108
Number of bits per angle	10
cyclic prefix	Long ie 32 samples
Number of $N_t \times N_r$ antennas	2×2 and 4×4
Modulation	QPSK and 64-QAM
FEC	Convolutional
Coding Rate	1/2

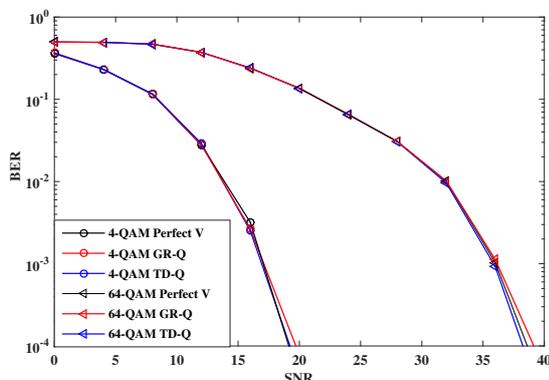


Figure 7: BER performance for perfect beamforming (Perfect V), beamforming with GR-Q and beamforming with the prposed time domain quantization (TD-Q) in MIMO 2×2 context. The ITU Channel Model is considered for simulations.

The BER performance is shown in Figures 7 and 8 for 4-QAM and 64-QAM in MIMO 2×2 and MIMO 4×4 context respectively. The number of bits per angle feedback is taken to 10. It can be seen that the proposed TD-Q presents the same performance, for all considered configurations in these simulations, than the GR-Q which is adopted in IEEE 802.11ac standard. As considered in the 802.11ac standard, the simulations confirm that 10 bits per angle feedback minimizes the quantization error and allows the quantized beamforming (GR-Q and TD-Q) to have the same performance than the perfect one.

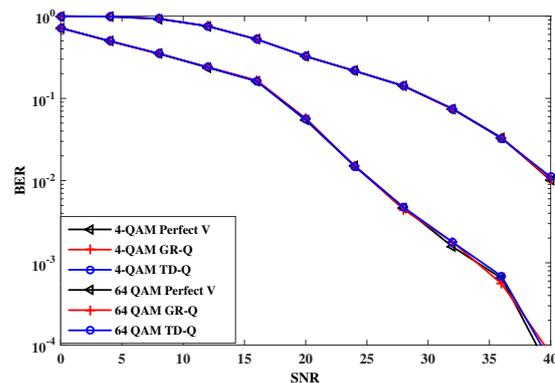


Figure 8: BER performance for perfect beamforming (Perfect V), beamforming with GR-Q and beamforming with the prposed time domain quantization (TD-Q) in MIMO 4×4 context. The ITU Channel Model is considered for simulations.

6 Conclusion

The proposed TD-Q requires less amount of feedback than the GR-Q one in IEEE 802.11ac TxBF mode. In addition, simulations in 802.11ac context with standard TGN channel model show that the TD-Q beamforming allows to have the same performance than the adopted GR-Q. The highlight of this proposal is that the proposed TD-Q can be considered as a credible alternative to the GR-Q for the IEEE 802.11ac evolution. The proposed quantization technic is not only for WLAN systems. It can be adapted for all SISO or MIMO system using closed loop beamforming.

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