

# Virtual Antenna Mapping MIMO Techniques in a Massive MIMO Test-bed for Backward Compatible LTE Mobile Systems

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**Abstract**—This paper proposes a virtual antenna mapping method for backward compatible massive or large-scale antenna multiple input multiple output (MIMO) base stations that provide communication services for legacy user equipment (UE) that can recognize only two or four base station antennas. The proposed method adopts and improves the omnidirectional beamforming that has been pioneered in previous works with proposing new method of determining antenna array coefficients through shifting the discrete Fourier transform (DFT) basis vectors for Zadoff-Chu (ZC) sequences. In the proposed method, the number of the parameters to be optimized is only two although the number of transmit antennas is hundreds or more (e.g., 500 antennas was proved in the paper as an example). Moreover, with the independent properties of the shifted versions of ZC sequences, this paper proves the fact that the coefficient vectors consisting virtual transmit antennas are independent when the channel gains work as the coefficients of them. This characteristic gives diversity with the pre-codes because two pre-code vectors must independent which means their linear combination with the non-zero channel gains or coefficients cannot be zero. The computer simulation results provide four important findings; the most important is that the actual number of virtually mapped physical antennas is inversely proportional to the transmit power per antenna.

**keyword**—massive MIMO, virtual antenna techniques, omnidirectional beamforming, transmit diversity, precoding techniques

## I. INTRODUCTION

Along with small cell technology, massive multiple input multiple output (MIMO) is a promising technology for increasing the capacity to fulfil the potential traffic increases in next-generation mobile communication systems [1]. While small cell technology can increase capacity through reducing the cell radius, massive MIMO adopts a large scale for the

transmit antennas. In small cell technology, in order to reduce the cell radius, the macro base station extends its antennas with some radio parts called ‘remote radio heads’ (RRHs) through connecting them with optical fibers, which results in high installation costs [1]. In contrast, by collocating many antennas to save the connecting costs massive MIMO technology has a relatively cheaper installation cost than the small cell technology [1]. Furthermore, massive MIMO has many radio frequency (RF) chains and antennas; thus, the transmission power per antenna decreases given that total transmission power is constant, which also results in decreasing RF part costs. Moreover, theoretically, massive MIMO technology can eliminate fast fading while increasing throughput and terminal numbers independent of the cell size [1-3]. Therefore, we designed massive MIMO test-bed as the pilot for next-generation mobile systems and future LTE-A system [4].

For a massive MIMO base station, dozens of or even hundreds of transmitting antennas are essential to give the effects described above. However, currently the LTE-A specification (e.g., Release-10 or later versions) allows eight transmit antennas and is expected to have more transmit antennas in the future [5, 6]. Therefore, just for analogy purpose, we first restrict the number of transmit antennas to eight in the paper although the massive-MIMO test-bed will have much more transmit antennas. However, with the simple example it is shown the possibility of extending them to hundreds by using the proposed method (e.g., 500 antennas in Figure 3b in this paper).

As for the backward compatibility, LTE-A eNode-B (i.e., base station) that has eight (or more) transmit antennas should provide service to legacy LTE (e.g., Release-8, 9) UEs that can recognize only two or four transmission antennas of eNode-B without notifying any information through additional control channels or signaling. Therefore, this paper introduces virtual antenna technique that can improve UE performance without increasing complexity of UE or eNode-B.

The proposed virtual antenna technique or backward compatible MIMO technique that converts eight physical transmit antenna to two or four logical antennas uses pre-codes that enable UE to use wireless channel’s degrees of freedom and diversity efficiently by having robustness for high spatial correlations. In order to give diversity with the pre-codes, two pre-code vectors must independent which means their linear

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combination with the non-zero channel gains or coefficients cannot be zero. The generation method for pre-coding matrix is similar but not same with cyclic delay diversity (CDD) that current LTE specification defines. The validation of the performance is proved by computer simulation because in the design stage, computer simulations can also be useful for good designing while test-beds are used in lab tests as a replacement for field tests.

In the following chapter, we introduce proposed virtual antenna mapping as the backward compatible MIMO techniques in massive MIMO test-bed with key technique of omnidirectional beamforming (OB), and we discuss about the transmitter and receiver structure for legacy user equipment (UE). Next, we evaluate the proposed method with some assumption for simple analysis. After evaluation, we conclude the paper with some important findings.

## II. PROPOSED VIRTUAL ANTENNA MAPPING

The proposed virtual antenna mapping that converts eight physical transmit antenna to two or four logical antennas is based on the OB as shown in Figure 1. With a OB beam, we can have two or four orthogonal beams by phase shifting of the beam precoding vector of the OB beam as we will show later.

### A. Omnidirectional Beamforming (OB)

The simplest way to make OB beam is sending transmit signal through only one antenna but this make the all transmit signal power converged into only one antenna and all power signals flow the radio frequency (RF) circuits of that antenna path only. As a result, high cost RF circuits like high power amplifier (HPA) must be needed [7]. Therefore, we can have following two requirements (RQs) for OB:

- RQ1. All radiating power for each transmit antenna in the MIMO BS must be equal or nearly equal.
- RQ 2. The gains of beam pattern have the least variations within the angle range of the service sector (e.g., 120 degree for 3 sector cell).

To find OB with above RQs, pioneering technique called random beamforming (RBF) has been used [7-9]. We make

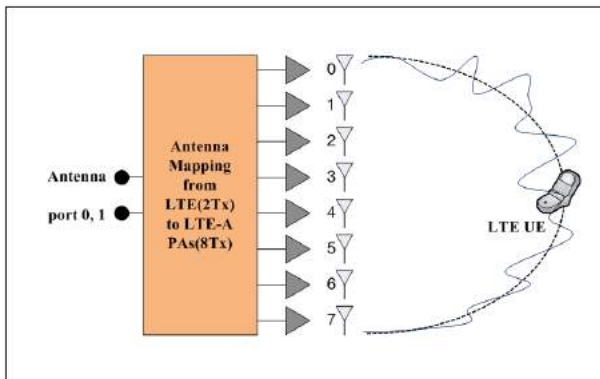


Fig. 1. Virtual antenna mapping concept for backward compatible MIMO techniques

RBF fancier technique by approaching the problem for satisfying above RQs with more systematic way of using the duality property of the digital Fourier transform (DFT) instead of using randomness which is described in next few paragraphs.

As the initial model, planer array antennas are commonly used and practical model can be develop based on this simple starting model. Therefore, for a simple analysis this paper assumes the system of using a uniform linear array (ULA) antenna and the beamforming array vector is given as follows [10]:

$$\bar{a}(\theta) = \left[ 1, e^{-j\frac{2\pi}{\lambda}d\sin(\theta)}, \dots, e^{-j\frac{2\pi}{\lambda}(M-1)d\sin(\theta)} \right]^T \quad (1)$$

where  $\lambda$  is the RF wave length,  $d$  is the distance between two adjacent antenna elements,  $\theta$  is the angle variable representing the direction of the beam pattern generated by the beam steering vector. Note that the analysis in reference [11] uses  $\phi$  the angle of incidence of the line-of-sight onto the transmit antenna array instead of using  $\theta$  (i.e.,  $\theta = \frac{\pi}{2} - \phi$ ) and so uses directional cosine (i.e.,  $\Omega := \cos(\phi)$ ) constructing the unit spatial signature for the angular domain analysis [11].

Suppose that  $\bar{w}$  is the weight coefficient vector with complex number entries  $\{w(m), m = 0, 1, \dots, M-1\}$  for the antenna elements where  $M$  is the total number of base station antennas. Then the resulting beam pattern transmitted by the total base station transmit antennas can be expressed as follows [10]:

$$G(\theta) = \bar{w}^H \cdot \bar{a}(\theta) = \sum_{m=0}^{M-1} w^*(m) \cdot e^{-j\frac{2\pi}{\lambda}md\sin(\theta)}. \quad (2)$$

Let's define a function  $\Omega(\theta) := \frac{2\pi}{\lambda}d\sin(\theta)$  and use this as the substitution variable. Note that reference [11] defines directional cosine (e.g.,  $\Omega := \cos(\phi)$ ) instead of  $\Omega(\theta)$  defined in this paper. Through the substitution, the beam pattern can be expressed as follows:

$$G(\theta) = W(\Omega(\theta)) = \sum_{m=0}^{N-1} w(m) \cdot e^{-j\Omega(\theta)m}$$

$$\text{or } W(\Omega) = \sum_{m=0}^{N-1} w(m) \cdot e^{-j\Omega m}. \quad (3)$$

This equation is a discrete-time Fourier transformation (DTFT) that transforms the discrete-time signal  $w(m)$  with a finite duration into a continuous and periodic frequency signal  $W(\Omega)$ . Therefore, using the inverse DTFT (IDTFT) for (3), we can obtain the sequence elements as follows:

$$w(m) = \frac{1}{2\pi} \int_{2\pi} W(\Omega) \cdot e^{j\Omega m} d\Omega. \quad (4)$$

As a result, our problem refers to finding the sequence  $\{w(m), m = 0, 1, \dots, M-1\}$  through simply performing the IDTFT of the continuous beam pattern  $W(\Omega)$  that was

originally from  $G(\theta)$  whose variance is minimal. Therefore, the sequence must be satisfied regardless of the domains that it is transformed to using (3) and (4) (i.e. DTFT and IDTFT), and the amplitude or power of each element of the sequence must be constant. The typical sequence with those characteristics is Zadoff-Chu (ZC) sequence and it is expressed as follows [12]:

$$s(m) = e^{-j\pi\mu m(m+\langle M \rangle_2)/M}, \quad (5)$$

where  $\mu$  is an integer less than and relative coprime to  $M$  and the notation  $\langle x \rangle_y$  denotes the remainder of  $x$  modulo  $y$ . Through inserting the circular shift parameter  $k$  into (5), we can obtain following equation:

$$\begin{aligned} \alpha(m) &:= s(m+k) \\ &= e^{-j\pi\mu\langle m+k \rangle_M} \langle m+k+\langle M \rangle_2 \rangle_M / M. \end{aligned} \quad (6)$$

With (6) and given the total number of base station transmit antenna elements  $M$ , the goal becomes narrowing this to locate only two parameters of  $\mu$  and  $k$  that can generate the unique circular-shifted ZC sequence with a large number of elements. Our goal is to make OB in angular domain which gets discrete coefficients for each antenna element through IDFT: so we can make resulting beam generated by the antennas in the original angular domain which will be the target constant amplitude beam again. Therefore, by considering beam steering vectors as the DFT operators, we choose the weighting sequence whose amplitude remains constant before and after DFT.

**B. Parameter Optimization Procedure**

In order to optimize parameter values for the antenna coefficient sequences, we first look at the relationships between antenna elements and beam patterns. Figure 2 shows the relationship of the sequence with the antenna weights  $\{w(m), m = 0, 1, \dots, M\}$ , approximated beam pattern  $W(\Omega)$ , and exact beam pattern  $G(\theta)$  (see upper parts of the figure). The antenna weights  $w(m)$  are transformed to approximated beam pattern  $W(\Omega)$  by DTFT, and turn further into exact beam pattern  $G(\theta)$  by the variable substitution process (see the upper part of the figure). In the process of the variable substitution and given the array length  $L$  for the array antennas, the sine domain or directional cosine domain (i.e.,  $\Omega$  domain) has uniform resolution ( $\frac{1}{2L}$ ) but the angular domain (i.e.,  $\phi$  domain) has non-uniform resolution [11]. With these relationships analysed in this paper, interesting readers can solve the exercise 7.8 in [11].

Figure 2 also shows the relationship between discrete Fourier transform (DFT) and DTFT to identify the sampling errors occurred in the process of converting signals between digital and continuous domains (see the lower box in the middle part of the figure). The sampling errors also give severe effects to increasing the variance of beam pattern and these errors can be decreased as the  $M$  increases, which are favorable to massive MIMO systems. With these relations, the idea of the proposed method come from the fact that Fourier

transform or its inverse transform of the ZC sequences can change the total average sequence amplitude and phases but can not change the amplitudes or powers of the sequence elements independently by the properties of the sequences as described in the bottom of the figure. For parameter optimizing procedures, up to date the most pioneered method has turned out to be the method of Yang, et. al.'s [7] which we follows in the point of variance minimizing target. Assume that for entire section  $\{\theta_1 \leq \theta \leq \theta_2 \mid \theta_1 = 0, \theta_2 = 2\pi\}$  and  $d = \lambda/2$  then the expected value of  $|G(\theta)|^2$  is given as  $M$  (i.e.,  $E[|G(\theta)|^2] = M$ ) [7]. Therefore, by using (2) and (4), the square variance can be calculated as follows (the detailed solving procedures are given in [7]).

$$\begin{aligned} \sigma^4 &= \frac{1}{2\pi} \int_0^{2\pi} (|G(\theta)|^2 - M)^2 d\theta \\ &= \frac{1}{2\pi} \int_0^{2\pi} \left( \left| \sum_{m=0}^{M-1} e^{-j\psi_m} \cdot e^{-j\pi m \sin(\theta)} \right|^2 - M \right)^2 d\theta \\ &= \sum_{j=1}^{M-1} \sum_{m=1}^{M-j} \sum_{n=1}^{M-j} \cos(\psi_{m+j} - \psi_m - (\psi_{n+j} - \psi_n)). \end{aligned} \quad (7)$$

Yang, et. al. obtained the optimum patterns by taking the partial derivatives with respect to  $\psi_m$  to reach the minimum value of  $\sigma^4$ . However, it is usually too complex to find the solution especially when the number of variables to find is several hundred with the common massive MIMO system nature.

Instead of using the partial derivative method, this paper uses relatively simple algorithm called the simplex search

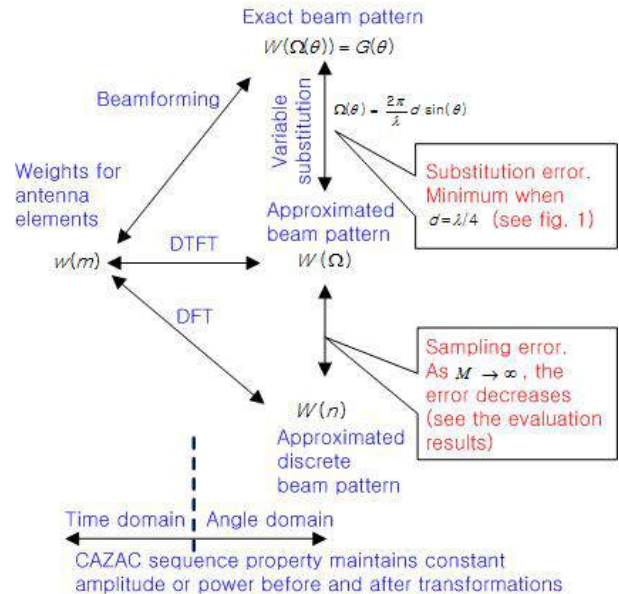


Fig. 2. The relationships for the antenna weights  $w(m)$ , approximated beam pattern  $W(\Omega)$ , and exact beam pattern  $G(\theta)$ . Important error sources of the sequences are also shown in two boxes located in right side.

method of Lagarias et. al. [13] which is a direct search method that does not use numerical or analytic gradients with expanding the method to n-dimensional space. While Yang’s method tried to optimize M variables for the antenna elements, the proposed method tries to optimize only two variables of the parameters  $\mu$  and  $k$  in (6) to make the sequence uniquely determined. Therefore, by substituting  $\alpha(m)$  in (6) to  $w(m)$  for (2) and (3), we can get following equation:

$$\sigma^4 = \sum_{j=1}^{M-1} \sum_{m=1}^{M-j} \sum_{n=1}^{M-j} \cos(\Psi_{m+j+k} - \Psi_{m+k} - (\Psi_{n+j+k} - \Psi_{n+k})) \tag{8}$$

where  $\Psi_x = -\pi\mu(\langle x \rangle_M)(\langle (x + \langle M \rangle_2) \rangle_M)/M$ .

While the beam pattern square variances  $\sigma^4$  in (7) and (8) look alike, the functions  $\Psi_x$  and  $\psi_m$  are defined differently. More precisely, the phase function of the sequence in (8) is more general form and that in (7) is specific form.

Figure 3a shows the polar plot for the found optimum antenna weights for 32 antenna elements  $\{w^*(m), m=0, 1, \dots, 31\}$ . These sequence is generated by the circular shifted Zadoff-Chu sequence rule with the optimized parameters of  $\{\mu^* = 1, k^* = 1\}$  which are founded by using the simplex search method with the initial value of  $\{\mu_o = 1, k_o = 1\}$  also for the proposed method. The computed values for them are  $\{\mu^* = 1.0157, k^* = 0.7366\}$  with the proposed method (see Figure 4). However  $\mu$  must be the relative coprime to the number 32 and  $k$  must be integer: so we choose them for the closest numbers that satisfy these conditions for the sequence. Figure 3b shows the same case with the Figure 3a except for the number of transmit antenna elements of 500.

The parameters for this sequence do not need to be found but predicted by the fact shown in the Figure 4 which shows the change trends for two optimized parameters  $\{\mu^*, k^*\}$  for the circular shifted Zadoff-Chu sequence as the number of transmit antenna increases. In the figure, they are converged into special value but we choose them for the closest numbers that satisfy these conditions for the sequence (i.e., relative coprime and integer).

To compare practical usefulness for the conventional method and proposed method in this paper, suppose the massive MIMO system that has 100 transmission antenna elements. Then, the number of variables to find in the proposed method decreases 50 times (i.e., 100:2) in these example procedures. Moreover, we found in the proposed method the optimum values of the two parameters of  $\mu$  and  $k$  changed with nearly negligible values as shown in above. Therefore, the proposed method can give good approximation information to the future analysis for the massive MIMO system with theoretically infinitive number of transmit antenna elements for (i.e.,  $M \rightarrow \infty$ ).

C. Virtual Antenna Mapping

Let DFT matrix be denoted by  $\underline{v} := [v_1, v_2, \dots, v_N]$  and  $\{v_i, i = 1, 2, \dots, N\}$  are the column vector set for the DFT

matrix. Further, suppose the first weighting sequence can be expressed as follows.

$$\bar{w}_0^{(1)} = \sum_{i=2}^{N-3} \alpha_i v_i \tag{9}$$

where  $\forall i, |\alpha_i| = 1$  for i-th complex scalar sequence to satisfy RQ1. In (9), we truncate the sequence or use only  $i=2, \dots, N-3$  instead of using all N elements of the sequence which means just only partial sequence elements can be used for the weights. Next, to fulfil RQ2 we can get the first OB weight  $\bar{w}_0^{(1)}$  by choosing and truncating the weighting sequence which satisfies (10) as follows.

$$\alpha_i^* = \underset{\{\alpha_i, i=m, \dots, n\}}{\operatorname{argmin}} \sqrt{\frac{1}{\theta_2 - \theta_1} \int_{\theta_1}^{\theta_2} (|G(\theta)|^2 - E[|G(\theta)|^2])^2 d\theta} \tag{10}$$

where  $G(\theta)$  is given by (2) and  $E[\cdot]$  denotes the expect value operator.

With an OB beam, eNode-B can map one virtual antenna port to multiple physical antennas (e.g., 8 transmit power

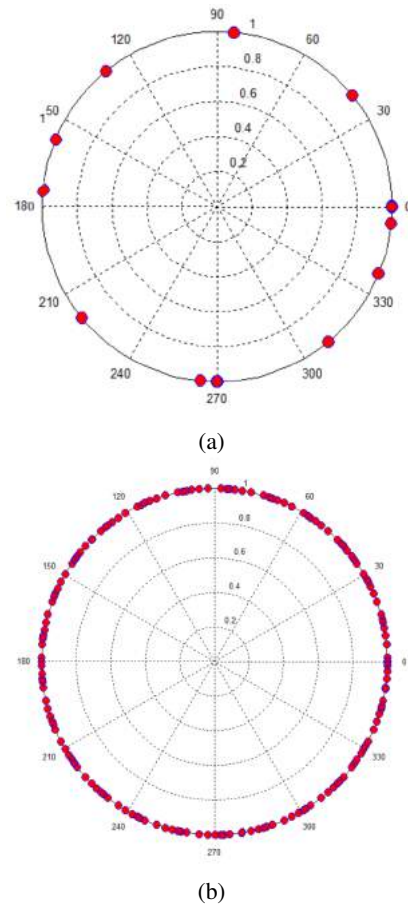


Fig. 3. Polar plot for the found optimum antenna weights for  $\{$  (a) 32, (b) 500  $\}$  antenna elements  $\{w^*(m), m=0, 1, \dots, 31$  or  $499\}$  by the proposed method. Some values are overlapped.

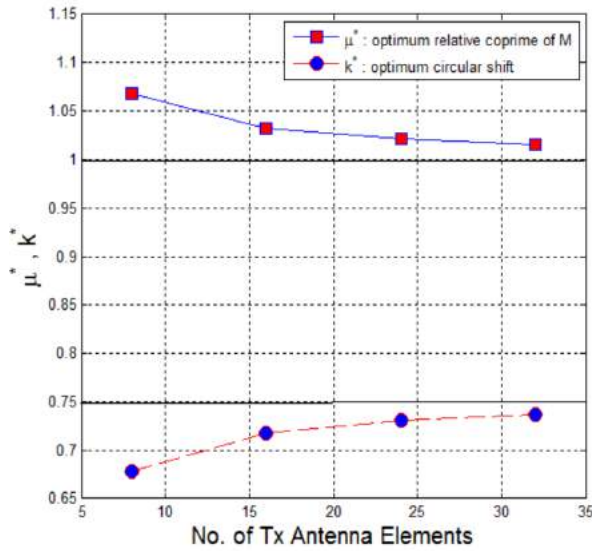


Fig. 4. Optimum circular shifted Zadoff-Chu sequence parameter value change as the number of transmit antenna elements increase for the proposed method

amp and antennas). To get one more virtual antenna port for diversity transmission, we need to get the second OB beam and this is given by  $\bar{w}_0^{(2)}$  as the circular shifted version of  $\bar{w}_0^{(1)}$  as follows.

$$\bar{w}_0^{(2)} = [\dots, w_{N-3}^{(1)}, w_{N-2}^{(1)}, w_{N-1}^{(1)}, w_0^{(1)}, w_1^{(1)}, \dots] \quad (11)$$

With the same manner, we can get third and fourth OB beams for four virtual antenna ports. However, we consider two OB beams for two virtual antenna ports and more than two OB beam case is left for future study.

In order to give diversity between two virtual transmit antennas, the coefficient vectors  $\bar{w}_0^{(1)}$  and  $\bar{w}_0^{(2)}$  must be independent when the channel gains work as the coefficients of them. Therefore, theorem 1 must be justified.

**Theorem 1.** Two vectors  $\bar{w}_0^{(1)}$  and  $\bar{w}_0^{(2)}$  generated by she shifted versions of ZC sequence are independent.

*Proof.* Suppose that  $\bar{w}_0^{(1)}$  and  $\bar{w}_0^{(2)}$  are dependent. Then there exists non-zero coefficients  $c_1, c_2$  such that  $c_1 \bar{w}_0^{(1)} + c_2 \bar{w}_0^{(2)} = 0$ . However,  $\bar{w}_0^{(1)}$  and  $\bar{w}_0^{(2)}$  are ZC sequences and with the auto-correlation property they are orthogonal each other and cannot be represented by linear combination of the other vector.  $\square$

With the manner described above, we can get sequences for beams 0 and 1 for eight physical antennas with power amplifiers (PAs) and example sequence values are shown in Table I. With the two OB beams of equations (9) and (11), the MIMO BS can map the two antenna ports to these two beams respectively. As an application example of the method for the traffic signals of the legacy UEs, cell-specific reference signals (CRSs) for diversity transmission with two antenna ports can be denoted by CRS0 and CRS1 and they can be directly mapped to OB beam 0 and 1 respectively.

With the same manner, the two physical broadcasting channel (PBCH) symbols for transmit diversity (e.g., Alamouti coded two symbols) can be mapped to OB beam 0 and 1 respectively.

TABLE I  
GENERATED BEAM WEIGHTING SEQUENCES

Beam 0 (e.g., $\bar{w}_0^{(1)}$ )	Beam 1 (e.g., $\bar{w}_0^{(2)}$ )
0.10406 + j*(-0.24340395)	-0.10406 + j*(+0.243403)
0.21346 + j*(-0.16386)	0.21346 + j*(-0.163868)
0.33082 + j*(+0.15179)	-0.33082 + j*(-0.151797)
-0.16340 + j*(+0.38933)	-0.16340 + j*(+0.389337)
-0.01444 + j*(-0.31956)	0.01444 + j*(+0.319562)
-0.08058 + j*(+0.40658)	-0.08058 + j*(+0.406583)
0.40698 + j*(+0.03329)	-0.40698 + j*(-0.033293)
0.20309 + j*(-0.25417)	0.20309 + j*(-0.254177)

D. Base Station’s Transmitter Structure for Legacy UEs

With the method described in previous chapter, we know how to get virtual antenna mapping. To find most efficient way to use the mapping for base station’s transmitter, we now consider two virtual antenna ports for the transmitter with eight physical antennas with following mapping cases:

- Case 1. Each port maps to two physical antennas with turning off the rest of the antenna PAs
- Case 2. Each port maps to two identical OB beams
- Case 3. Each port maps to two orthogonal OB beams with open loop space time block code (STBC)
- Case 4. Each port maps to two orthogonal OB beams with closed loop feedback for codebook index

Case 1 violates RQ1. and Case 2 has no diversity gain for two identical beams mapped by two virtual antenna ports, which will be proved at the evaluation chapter. On the other hand, Case 3 has diversity gain for two orthogonal beams mapped by two virtual antenna ports and used for STBC encoding as shown in Figure 5. This fact will be also proved at the evaluation chapter. For Case 4 we consider the transmitter and receiver mechanism shown in Figure 6. For the Case 4, we used LTE codebook index as shown in Table II [5].

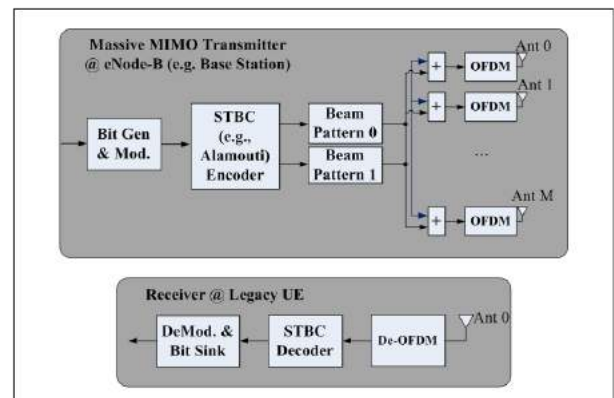


Fig. 5. Transmitter and receiver structure for Case 3

The codebook indices zero and one have zero valued element which means turning off the counterpart virtual antenna OB beam. The other codebook indices except those two show power distribute equally for the two beams which means equally combining of the two beams. The element's negative sign and imaginary component are intended to work for the signal phase change; so we can easily predict Case 3 and 4 will show the double (3 dB) performance difference which proved by the computer simulation results shown at the evaluation chapter.

TABLE II  
CODEBOOK FOR CASE 4

Id	0	1	2	3	4	5
VI	1 0	0 1	$s$ $s$	$s$ $\bar{s}$	$s$ $js$	$s$ $j\bar{s}$

$$*s = \frac{1}{\sqrt{2}}, \bar{s} = -s$$

III. EVALUATION

Proposed virtual mapping method for backward compatible MIMO is evaluated by the computer simulations.

A. Computer Simulation Setup and Method

The computer simulation parameters are set as shown in Table III. Transmitters at eNode-B and UE receiver are set according to the Cases 1 - 4 described in above section.

TABLE III  
SIMULATION PARAMETERS

Parameter	Values
Antenna configuration	8 elements, ULA, $d = \lambda/2$
Modulation	QPSK
Channel model	Random fading (not change for one frame duration and uncorrelated between antenna paths)
Packet length	4x130 bits/packet (260 symbols/packet or QPSK)
CSI feedback period (for Case4)	1 frame

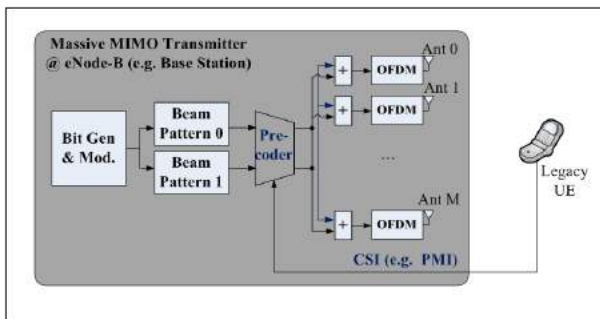


Fig. 6. Transmitter and receiver structure for Case 4.

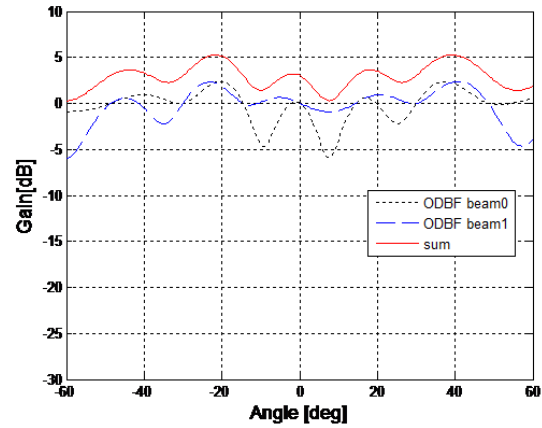


Fig. 7. Gains and their sum of the two OB beams

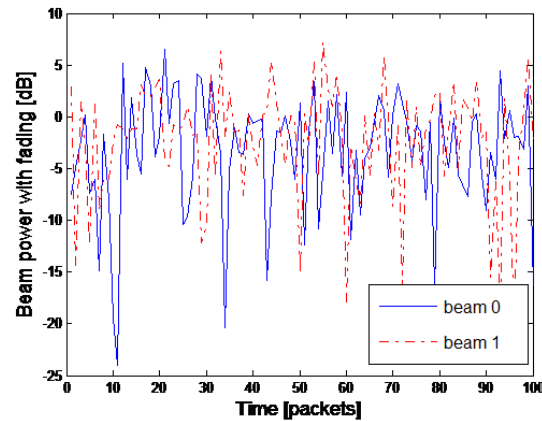


Fig. 8. Beam power variations with time for two orthogonal beams

B. Simulation Results and Discussion

Figure 7 shows the gains and their sum of the two draft OB beams obtained by the equations (9) and (11) that have example weighting sequences shown in Table 1. Approximately, each beam 0 and 1 fluctuates with 5 dB as the figure shows. However, we can get more flat OB beams by tuning the parameters of the equations but we leave this future study.

Figure 8 shows beam power variations with time for two orthogonal beams. In the figure, the power difference between the two beams shows distinctively and so gives diversity gain which can be seen in the Figure 9.

Figure 9 shows uncoded bit error rate (BER) performance versus transmit power per antenna for QPSK (a) and 16QAM (b). These figures as the simulation results give us following four findings:

- i) Approximately, the uncoded BER results for the Case 3 and Case 4 are 6 dB (4 times) more than that of Case 1.
- ii) Approximately, the uncoded BER results for the Case 4 are 3 dB (2 times) more than that of Case 3.
- iii) The curve for Case 1 is steeper than that for Case 2.
- iv) The uncoded BER performance results for QPSK and

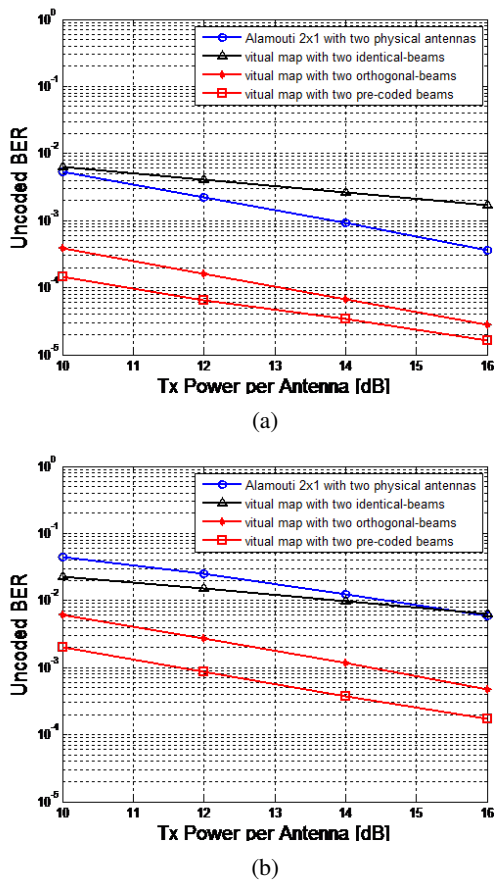


Fig. 9. Uncoded bit error rate performance versus transmit power per antenna: (a) QPSK, (b) 16QAM

16QAM show the same trend but the power difference shows 6 dB approximately.

The reason of i) is that with the same total powers of all transmitting physical antenna for each case the power per antenna is different. That is, Case 3 and Case 4 use eight transmit antennas while Case 1 uses two which result is four times (6 dB) power difference. The reason for ii) is that codebook index feedback of the Case 4 results in combining diversity gain and the phase control of the two beams. For example, the index (Id) 0 and 1 in Table 1 have the values (VI)  $[1, 0]^T$  and  $[0, 1]^T$  respectively which mean full transmit power allocation for the good transmission antenna path while no power allocation for the bad. Therefore, the BER performance becomes as double as the open loop transmit diversity (Case 3). For the fact of iii), Case 2's two identical beams show no diversity gain as Case 1 of STBC encoding which result in steeper uncoded BER curve. From the iv), we can predict high order modulation (e.g., 64QAM) can work with the same trend and can be adopted in the proposed scheme.

#### IV. CONCLUSION

This paper proposed a virtual antenna mapping method for backward compatible massive MIMO base stations in order to provide communication services for legacy user equipment

that can recognize only two or four antennas. The proposed method adopts the omnidirectional beamforming that satisfies the two research questions defined in this paper, and it provides a more systematic approach than previous pioneering works, as discussed in the introduction. The proposed method is to determine antenna array coefficients through shifting the discrete Fourier transform (DFT) basis vectors for Zadoff-Chu (ZC) sequences. With this method, this paper showed the possibility of extending the number of transmit antenna to hundreds (e.g., 500 antennas in Figure 3b in this paper) with only two parameters to be optimized. Moreover, with the independent properties of the shifted versions of ZC sequences, this paper proved the fact that the coefficient vectors consisting virtual transmit antennas are independent when the channel gains work as the coefficients of them. This characteristic enables the communication link having diversity with the two or more virtual transmit antennas. That is, in order to give diversity with the pre-codes, two pre-code vectors must independent which means their linear combination with the non-zero channel gains or coefficients cannot be zero. This paper also provides four types of transmitting structures at the base station in order to demonstrate the performance of the proposed virtual antenna mapping, and it derives four findings through a simple computer simulation (see Chapter III-B). Among these findings, the most important is that a four times (6 dB) higher uncoded BER performance was achieved when mapping eight physical antennas compared with that of mapping two physical antennas. Therefore, the actual number of mapped physical antennas is inversely proportional to the transmit power per antenna.

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