Chapter 3

Methodology

3.1 Introduction:

Multi-antenna wireless communication systems have demonstrated significant performance enhancement in recent years. Array technology, which is combined with space-time coding technique, is the biggest characteristic of space-time coding, Alamouti proposed a kind of transmit diversity scheme based on two transmit antennas. It can achieve full diversity and full rate by using simple Maximum Likelihood decoding algorithm at the decoder, but when transmitting antenna number is more than two, the space-time block code system which is based on the complex orthogonal design is unable to achieve full diversity and full rate. To solve this problem, a designed scheme of quasi-orthogonal space-time block codes that can achieve full rate is presented by Jafarkhani, but it is at the expense of the diversity gain of the system. Recently, improved quasi- orthogonal coding schemes by used beamforming technology, it achieved the purpose of improving to decrease the interference, the rotation of the constellation graph is utilized to increase the distance between the code words so that the bit error rate is reduced at the same time. Then a beamforming scheme combined with QOSTBC is used four independent antenna arrays to achieve full diversity .

This chapter is summarization of the work done in this study starting with investigating the channel capacity of MIMO system and its relative calculations, then demonstrating the system model used for the performance evaluation from the previously proposed and proven approaches using the optimum choices available for the QOSTBC scheme to be compared with the proposed solution which uses the beamforming technique in order to reduce the interference between the redundancy signals which led to better BER records.

3.2 MIMO channel capacity:

For a MIMO system with N_T transmit and N_R receive antennas, as shown in Figure 1 below, a narrowband time-invariant wireless channel can be represented by $N_R \times N_T$ deterministic matrix $H \in \mathbb{C}^{N_R \times N_T}$.

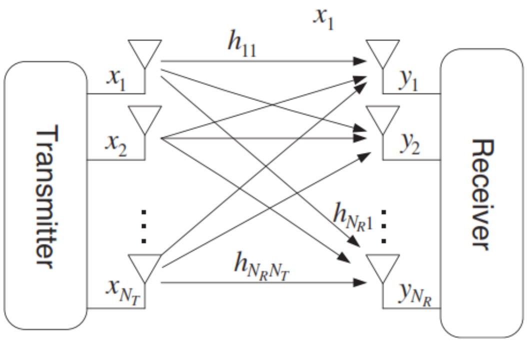

Figure (3.1): $N_R \times N_T$ MIMO system

Consider a transmitted symbol vector $x \in \mathbb{C}^{N_T \times 1}$, which is composed of N_T independent input symbol $x_1, x_2, \ldots, x_{N_T}$. Then, the received signal $x \in \mathbb{C}^{N_T \times 1}$ can be rewritten in a matrix form as follows:

$$y = \sqrt{\frac{E_x}{N_T}} Hx + Z \qquad (3.1)$$

Where $Z = (z_1, z_2, \ldots, z_{N_R})^T \in \mathbb{C}^{N_R \times 1}$, is the noise vector . This is assumed to be Zero-Mean Circular Symmetric Complex Gaussian (ZMCSCG). The noise vector Z is referred to as circular symmetric when $e^{j\theta} Z$ has the same same distribution as Z for any θ. Hence that there is no any channel information available at the transmitter side, one can spread the energy equally among all the transmit antennas, that is, the autocorrelation function of the transmit signal vector x is given as:

$$R_{xx} = I_{N_T} \quad (3.2)$$

Where I_{N_T} is a $N_T \times N_T$ identity matrix. Then the channel capacity is given as:

$$ChC = \log_2 det\left(T_{N_R} + \frac{E_x}{N_T N_0} HH^H\right) \quad (3.3)$$

Where E_x is the channel SNR at each receiving antenna, so the channel capacity for multiple input single output MISO is reduced and given as:

$$ChC_{MISO} = \log_2\left(1 + \frac{E_x}{N_0}\right) \quad (3.4)$$

Where

ChC is channel capacity for Multiple input Single output

Hence that a MIMO channel is converted into r virtual SISO channels with the transmit power E_x/N_T for each channel and the channel gain of γ_i for the i^{th} SISO channel. And we have assumed that MIMO channels are deterministic. In general, however, MIMO channels change randomly. Therefore, H is a random matrix, which means that its channel capacity is also randomly time-varying. In other words, the MIMO channel capacity can be given by its time average. In practice, we assume that the random channel is an ergodic[1] process. Then the channel capacity of the open loop MIMO system can be shown as:

$$ChC_{MIMO} = E\left\{\sum_{i=1}^{r} \log_2\left(1 + \frac{E_x}{N_T N_0}\gamma_i\right)\right\} \quad (3.5)$$

A comparison of ergodic channel capacity for different antenna configuration in terms of throughput versus SNR using the equations above is discussed later with results.

[1] A random process is ergodic if its time average converges to the same limit for almost all realizations of the process.

3.3 Space Time Block Codes STBC:

STBC generally has two categorizes; orthogonal and non-orthogonal STBCs, orthogonal STBC like Alamouti's STBC, non-orthogonal like Quasi STBC, Quasi STBC can be open loop or closed loop depending on a feedback returned to the transmitter or not, but when sending a redundant copies of signal from multiple number of antennas that equals to that number of copies it's unnecessary for that to fed the transmitter back with any acknowledgment about a different copies of a signal that it can decode, except in systems that has higher accurate requirements. An effective technique that can be used and is already deployed in different transmission issues is the beamforming, which will be combined with QOSTBC and QPSK modulation and expressed in this chapter.

3.4 STBC general structure:

The general structure of STBC transmit diversity used to send multiple copies of a signal from multiple antennas, by sending a time slot once a time but from different antennas, let a signal $S(t)$ consist of a sequence of symbols $x_1, x_2, x_3, \dots, x_n$ an let there is P antennas send these sequence, so there will be N_{seq}sequences to be sent from P_M antennas where, $N_{seq} = P_M$, where $M$is the maximum number of antennas, the STBC will generate $N_{seq}\ copies\ from\ S(t)$ with a proper correlation, this generated signal copies can be represented with the co-product relation as follow:

$$\mathrm{C}(t) = \coprod_{n=0}^{N_{seq}} S(t) \tag{3.6}$$

Where $C(t)$ is the transmit block which contains the different correlated coded sequences.

3.5 Real Orthogonal Designs:

The first STBC orthogonal design for real values as introduces in [28] states that, a real orthogonal design of size k is an orthogonal $k \times k$ matrix with entries the indeterminate $\pm x_1, \pm x_2, \pm x_3, \dots, \pm x_k$. The existence problem for orthogonal designs is known as the Hurwitz–Radon problem in the mathematics literature [29], and was completely settled by Radon in another context at the beginning of the previous century. In fact, an orthogonal design exists if and only if $k = 2, 4, or\ 8$. Given an orthogonal design C, one can negate certain columns of C to arrive at another orthogonal design where all the entries of the first row have positive signs. By permuting the columns, we can make sure that the first row of C is $x_1, x_2, x_3, \dots, x_k$. Thus we may assume without loss of generality that C has this property. Examples of orthogonal designs are the 2×2 design:

$$\begin{bmatrix} x_1 & x_2 \\ -x_2 & x_1 \end{bmatrix} \tag{3.7}$$

For the 4×4 design let consider the previous 2×2 is A, and add another 2×2design with x_3, x_4, to be B as:

$$B = \begin{bmatrix} x_3 & x_4 \\ -x_4 & x_3 \end{bmatrix} \tag{3.8}$$

Then the 4×4 design will be:

$$\begin{bmatrix} \mathrm{A} & \mathrm{B} \\ -\mathrm{B}^T & \mathrm{A}^T \end{bmatrix} = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \\ -x_2 & x_1 & -x_4 & x_3 \\ -x_3 & x_4 & x_1 & -x_2 \\ -x_4 & -x_3 & x_2 & x_1 \end{bmatrix} \tag{3.9}$$

Where $-\mathrm{B}^T$ represents the negative transpose of the matrix B. on the other hand, for the 8×8 design using the same technique, by using the 4×4 design as A, and add another 4×4 design with x_5, x_6, x_7, x_8 to be B as:

$$B = \begin{bmatrix} x_5 & x_6 & x_7 & x_8 \\ x_6 & -x_5 & -x_8 & x_7 \\ x_7 & x_8 & -x_5 & -x_6 \\ x_8 & -x_7 & x_6 & -x_5 \end{bmatrix} \quad (3.10)$$

Then the 8×8 design will be:

$$\begin{bmatrix} A & B \\ -B^T & A \end{bmatrix} = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 & x_5 & x_6 & x_7 & x_8 \\ -x_2 & x_1 & x_4 & x_3 & x_6 & -x_5 & -x_8 & x_7 \\ -x_3 & -x_4 & x_1 & x_2 & x_7 & x_8 & -x_5 & -x_6 \\ -x_4 & x_3 & -x_2 & x_1 & x_8 & -x_7 & x_6 & -x_5 \\ -x_5 & -x_6 & -x_7 & -x_8 & x_1 & x_2 & x_3 & x_4 \\ -x_6 & x_5 & -x_8 & x_7 & -x_2 & x_1 & -x_4 & x_3 \\ -x_7 & x_8 & x_5 & -x_6 & -x_3 & x_4 & x_1 & -x_2 \\ -x_8 & -x_7 & x_6 & x_5 & -x_4 & -x_3 & x_2 & x_1 \end{bmatrix} \quad (3.11)$$

Then to define the complex orthogonal STBC is invented as, the matrices of 2×2 and 4×4 designs can be identified, respectively, with complex number $x_1 + x_2 i$ for the 2×2 design, and the quaternion number $x_1 + x_2 i + x_3 j + x_4 k$ for the 4×4 design. And this was a leading factor for the existence of the Alamouti's and Quasi STBCs and other orthogonal STBC designs.

3.6 Orthogonal Designs:

Orthogonality is the relation of two lines at right angles to one another (perpendicularity), and the generalization of this relation into *n* dimensions; and to a variety of mathematical relations thought of as describing non-overlapping, uncorrelated, or independent objects of some kind [29].

STBCs as originally introduced, and as usually studied, are orthogonal. This means that the STBC is designed such that, "the vectors representing any pair of columns taken from the coding matrix is orthogonal". The result of this is simple, linear, optimal decoding at the receiver. Its most serious disadvantage is that all but one of the codes that satisfy this criterion must sacrifice some proportion of their data rate (as in Alamouti's code).

An orthogonal design of complex variables $x_1, x_2, x_3, \ldots, x_n$ is a $p \times k$ matrix $C(x)$which denotes the following:

- The entries of $C(x)$ are complex linear combination such like $[x_1, x_1^*, x_2, x_2^*, , \ldots, x_n, x_n^*]$.
- $C^H C = (|x_1|^2 + |x_2|^2 + \cdots + |x_n|^2) I_k$, where I_k is an identity matrix of size $k \times$ k , and the rate of $C(x)$ is defined as: $\left({}^{n}/_{p}\right)$.

3.7 Alamouti's STBC:

Figure (1) below shows the simplest transmit diversity scheme should have two antennas , then as assumed in Alamouti's STBC there should be P_1, P_2 antennas and two conjugate correlated copies of the signal sequence , So each conjugate correlated copy of the sequence will be sent symbol by symbol as the correlation dependence.

Hence, $S(t) = \{x_1, x_2, x_3, \ldots, x_n\}$, then the correlation dependence between the original sequence $S(t)$, and it copies are combined as $\mathrm{C}(x)_L$ where L is the number of copies or layers which is identical to the number of antennas P_M and as assumed here equals two so $N_{seq} = 2$, then $\mathrm{S}(t)_1, \mathrm{S}(t)_2$ are described as:

$$\mathrm{S}(t)_1 = \{x_1, x_2\} \quad (3.12)$$

$$\mathrm{S}(t)_2 = \{-x_2^*, x_1^*\ \} \quad (3.13)$$

Where $\mathrm{S}(t)_1$ and $\mathrm{S}(t)_2$ are the original sequence and the correlated sequences that will be sent from different antennas, 1 and 2 represents the antennas number [30]. Then the full diversity and full code rate transmit block codes orthogonal design defined by Alamouti's 2x2 STBC is defined as follow :

$$C(x)_2 = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix} \tag{3.14}$$

Where the number of columns represents the number of antennas and the number of rows represent signal sequence length or time-steps, this design is obtained from the real orthogonal STBC 2×2 design defined before. Hence that the signal power will be split between two antennas there will be two time-steps for full rate diversity transmission.

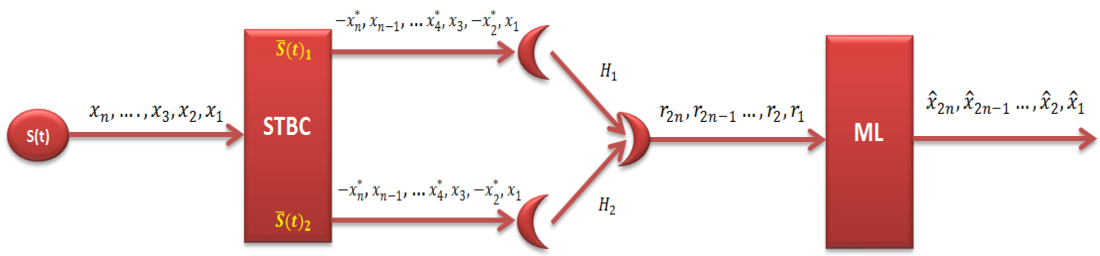

Figure (3.2) :optimal Alamouti's 2x1 transmit diversity scheme with one receiver

For modulated symbol coding using Alamouti's STBC with AWGN at receiver, the transmitted symbols x_1, x_2 are formed as s_1, s_2 as shown in figure (3.3) below:

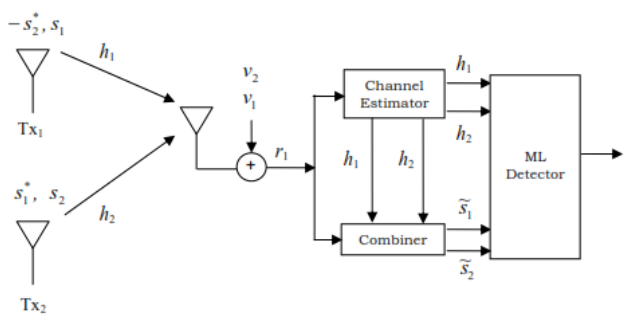

Figure (3.3): Alamouti's 2x1 transmit diversity scheme with one receiver

Here s_1, s_2 represents the modulated transmit symbols, h_1, h_2 are the fading channel factors, v_1, v_2 represent the receiver noise that will be added to the received symbols.

For three and four transmit antennas, there are codes of rate 3/4 [31], defined as:

$$\mathrm{C(x)}_3 = \begin{bmatrix} x_1 & x_2 & x_3 \\ -x_2^* & x_1^* & 0 \\ -x_3^* & 0 & x_1^* \\ 0 & -x_3^* & x_2^* \end{bmatrix}, \mathrm{C(x)}_4 = \begin{bmatrix} x_1 & x_2 & x_3 & 0 \\ -x_2^* & x_1^* & 0 & x_3 \\ -x_3^* & 0 & x_1^* & -x_2 \\ 0 & -x_3^* & x_2^* & x_1 \end{bmatrix} \qquad (3.15)$$

3.8 Quasi Orthogonal STBC:

The first proposed quasi orthogonal designs for STBC are introduced by Papadias, Foschini [32] & Jafarkhani [33]. The Quasi OSTBC defined a full rate orthogonal design for four transmit antenna diversity using the same methodology used in real orthogonal STBC designs, by using the Alamouti's block as A and defines the same block for x_3, x_4 as:

$$A = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix}, \qquad B = \begin{bmatrix} x_3 & x_4 \\ -x_4^* & x_3^* \end{bmatrix} \qquad (3.16)$$

Then the QOSTBC defined in [34] which constructed from Alamouti's STBC, can be shown as:

$$\mathrm{C} = \begin{bmatrix} \mathrm{A} & \mathrm{B} \\ -\mathrm{B}^* & \mathrm{A}^* \end{bmatrix} = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \\ -x_2^* & x_1^* & -x_4^* & x_3^* \\ -x_3^* & -x_4^* & x_1^* & x_2^* \\ x_4 & -x_3 & -x_2 & x_1 \end{bmatrix} \qquad (3.17)$$

Moreover, there exist quasi-orthogonal STBCs that achieve higher data rates at the cost of inter-symbol interference (ISI). Thus, their error-rate performance is lower bounded by the one of orthogonal rate 1 STBCs that provides ISI free transmissions due to orthogonality, and then the code above does not have full diversity as the STBCs theorems states.

$$C^H C = \begin{bmatrix} a & 0 & 0 & b \\ 0 & a & -b & 0 \\ 0 & -b & a & 0 \\ b & 0 & 0 & a \end{bmatrix} \quad (3.18)$$

Where $(\cdot)^H$denotes the Hermitian transpose, and we can say that a matrix is Hermitian if it equal its own conjugate transpose and its square matrix and its elements are complex. Then (a) and (b) can be estimated by:

$$a = |x_1|^2 + |x_2|^2 + |x_3|^2 + |x_4|^2 \quad (3.19)$$

$$b = x_1 x_4^* + x_4 x_1^* - x_2 x_3^* - x_3 x_2^* \quad (3.20)$$

The use of ML decoding requires two cost functions for this scheme, so these two cost functions should be $f_1(s_1, s_4)$ and $f_2(s_2, s_3)$, then the ML decision metric of this code can be written as the sum of two terms $f_1(s_1, s_4\ ) + f_2(s_2, s_3)$, where f_1 depends only on s_1 and s_4 , and f_2 depends only on s_2 and s_3 . Thus, the minimization can be done separately on these two terms, i.e., symbol pairs (s_1, s_4) and (s_2, s_3) can be decoded separately, which leads to fast ML decoding.

From (3.20), the minimum rank of the difference matrix between two distinct codewords is 2, which means that the code in (3.17) does not have full diversity. This is why we use the rotation technique as it h as been proven in [35] that the full diversity can be achieved for this code with the rotation technique.

Another code is also proposed with the code (3.19) in [36].but with the following structure:

$$\begin{bmatrix} A & B \\ -B & A \end{bmatrix} \tag{3.21}$$

In this code the problem is that, the modulated transmitted symbols S_1, S_2, S_3, S_4 cannot be separated into two groups at the receiver for the fast ML decoding.

Another similar scheme (called TBH) for quasi-orthogonal STBCs also has been proposed in [37]. For four transmit antennas, the TBH scheme is:

$$C = \begin{bmatrix} A & B \\ B & A \end{bmatrix} = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \\ -x_2^* & x_1^* & -x_4^* & x_3^* \\ x_3 & x_4 & x_1 & x_2 \\ -x_4^* & x_3^* & -x_2^* & x_1^* \end{bmatrix} \tag{3.22}$$

Where A and B are similar as in (3.18), then:

$$C^H C = \begin{bmatrix} a & 0 & b & 0 \\ 0 & a & 0 & b \\ b & 0 & a & 0 \\ 0 & b & 0 & a \end{bmatrix} \tag{3.23}$$

Where a is similar as in (3.19), and:

$$b = x_1 x_3^* + x_3 x_1^* - x_2 x_4^* - x_4 x_2^* \tag{3.24}$$

3.9 Rotated Quasi STBC:

Sometimes, it is impossible to achieve code rate 1 for the complex orthogonal codes. To provide full diversity, different constellations are sending through different transmitted symbols. This is done by rotating the symbols before transmission. This provides full-diversity with code rate 1 and this pairing of symbols gives good performance as compared to QOSTBC.

For M receive antennas, a diversity of 1*M and 4M* is achieved while the rate of the code is one. The maximum diversity of 4*M* for a rate one complex orthogonal code is impossible in this case if all symbols are chosen from the same constellation. By using same constellation for all symbols in the subset reduces the minimum distance for such codes. As a remedy to this problem, rotation based method is used that aims in maximizing the minimum distance in the space time constellation. To provide full diversity, we use different constellations for different transmitted symbols.

For example, we may rotate symbols x_3 and x_4 before transmission. Let us denote $\overleftrightarrow{x_3}, \overleftrightarrow{x_4}$ as the rotated versions of $x_3\ and\ x_4$, respectively. It's proven that it is possible to provide full-diversity QOSTBCs by replacing $x_3\ and\ x_4$ with $\overleftrightarrow{x_3}\ and\ \overleftrightarrow{x_4}$.The resulting code is very powerful since it provides full diversity, rate one, and simple pair wise decoding with good performance. Different modulation techniques use different rotation. In this study, we are using QPSK as shown in figure (3.4) and 16QAM as in figure (3.6) .The optimum rotation for QPSK constellations used is $\pi/2$ as resulted in figure (3.5) .

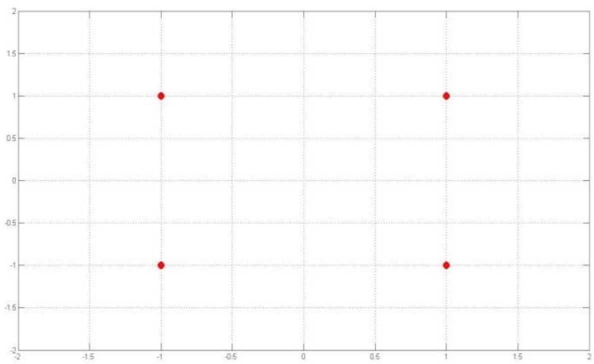

Figure (3.4): QPSK Constellation without rotation

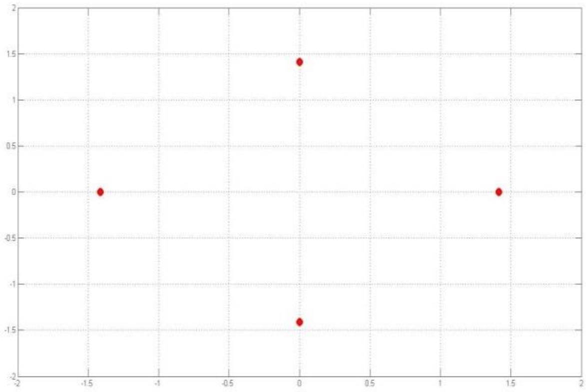

Figure (3.5): QPSK constellation with rotation

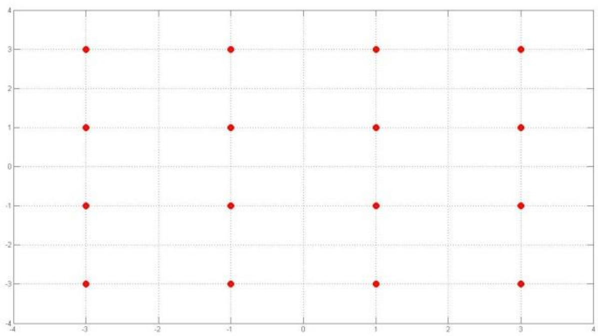

Figure (3.6):16QAM constellation

3.10 Additive White Gaussian Noise (AWGN) :

The additive white Gaussian noise (AWGN) provides a simplistic view of a communication channel by modelling the presence of noise without accounting for distortion due to signal fading.

The AWGN [38] channel is a discrete-time channel with continuous input and output variables. At time t , the channel output Y_t is defined as:

$$Y_t = S_t + n_t \tag{3.25}$$

Where S_t is the transmitted signal at independent of n_t , which is the additive noise term variance with the Gaussian distribution, and its equivalent covariance matrix.

3.11 Beamforming:

The beamforming technique was originated from array processing, it utilizes the available antenna elements and signal processing algorithms to adjust the strength of the transmitted and received signals based on their departure or arrival directions. In wireless communication applications, the “direction” can be either the physical direction for phase array in line-of-sight (LOS) environment, or the direction in a mathematical sense for the matched channel information and the focus of energy is achieved by choosing appropriate weights for each antenna element under a certain criterion.

Beamforming is a well-known complementary technique in communication system by provides array gain. MIMO systems which exploit beamforming and space-time block coding (STBC) have recently attracted enormous interest due to its potential to enhance performance and increase the capacity of mobile communication system[39]. The beam propagation distribution function of beamforming can be described as:

$$\rho(D) = \frac{1}{D}\sum_{k=1}^{K} \exp(j\pi \sin(\theta_0 + \Delta\theta_k)) \qquad (3.26)$$

Where D is the separation of antenna elements in wavelength; K is the number of sub-paths; θ_0 is the mean Angle of Arrival AOA or Angle of Departure AOD, and $\Delta\theta_k$ is the k^{th} offset angle in radians.

3.11.1 Transmit Beamforming (TxBF):

The implementation of adaptive antenna array technique in a handset is difficult with today`s hardware due to its limitations in size, cost, and energy storage capability, while it is feasible to adopt antenna arrays at base stations.

Transmit beamforming provides apowerful method for increasing downlink capacity. The idea of TxBF is similar to the pre-coded MIMO technique but with different strategies to calculate the transmit weight vector. TxBF adjusts the antenna main lobe towards to the desired user and reduce the interference to other users.

3.11.2 Receive Beamforming (Rx BF):

Beamforming also can be applied in the uplink to improve the link quality and suppress the co-channel interference, which is known as receive beamforming (RxBF). Through RxBF, smart antenna system can receive predominantly from a desired direction (direction of the desired source) compared to some undesired directions (direction of interfering sources). This implies that the digital processing has the ability to shape the radiation pattern to adaptively steer beams in the direction of the desired signals and put nulls in the direction of the interfering signals. This enable low co-channel interference and large antenna gain to the desired signal.

3.12 Maximum Likelihood decoding (ML):

The proposed system model used in this study is uses the QOSTBC scheme mentioned in (3.8).So after the signal transmitted passing through the fading channel, the faded symbols are received at the receiver antenna with the additive noise described before. For QOSTBC system in use, we use pair wise maximum likelihood decoding. We can decode a pair of symbols independently in QOSTBC. We can decode the received symbols using the following equations :

The pair wise maximum likelihood cost function for $s_1 and\ s_4$:

$$f_{1_4}(s_1, s_4) = \sum_{m=1}^{M} \Bigg[(|s_1|^2 + |s_4|^2) \left(\sum_{n=1}^{4} |a_{n,m}|^2 \right) + 2\Re\{ (-a_{1,m} r_{1,m}^* - a_{2,m}^* r_{2,m} - a_{3,m}^* r_{3,m} - a_{4,m} r_{4,m}^*) s_1 + (-a_{4,m} r_{1,m}^* + a_{3,m}^* r_{2,m} + a_{2,m}^* r_{3,m} - a_{1,m} r_{4,m}^*) s_4 \} + 4\Re\{ a_{1,m} a_{4,m}^* - a_{2,m}^* a_{3,m} \} \Re\{ s_1 s_4^* \} \Bigg] \quad (3.17)$$

The pair wise maximum likelihood cost function for $s_2 and\ s_3$:

$$f_{2\_3}(s_2, s_3) = \sum_{m=1}^{M} \Bigg[(|s_2|^2 + |s_3|^2) \left(\sum_{n=1}^{4} |a_{n,m}|^2 \right) + 2\Re\{ (-a_{2,m} r_{1,m}^* + a_{1,m}^* r_{2,m} - a_{4,m}^* r_{3,m} + a_{3,m} r_{4,m}^*) s_2 + (-a_{3,m} r_{1,m}^* - a_{4,m}^* r_{2,m} + a_{1,m}^* r_{3,m} + a_{2,m} r_{4,m}^*) s_3 \} + 4\Re\{ a_{2,m} a_{3,m}^* - a_{1,m}^* a_{4,m} \} \Re\{ s_2 s_3^* \} \Bigg] \quad (3.18)$$

Where s_1, s_2, s_3 and s_4 are the transmitted symbols, and a is the path matrix, and r is the symbol matrix.

3.13 Bit Error Rat

After the simulation script is executed with the described processes the performance evaluation factor BER is estimated and recorded for the comparison. The bit error rate represents the statistical proportion of bits received in error. The BER is expressed as [40]:

$$BER = \frac{1}{M} \sum_{u=1}^{M} P_b(u) \quad (3.27)$$

Where:

$$P_b(u) = \frac{1}{\log_2 M}\sum_{j=1}^{M} D_H(j,u)\cdot P(\hat{S} = S_j | S = S_u)$$

$$= \frac{1}{\log_2 M}\sum_{j=1}^{M} D_H(j,u)\left(1 - P_u\left(\bigcup_{i\neq j} \varepsilon_{ji}\right)\right) \qquad (3.28)$$

Where $m = \log_2 M$ is the number of data bits and $D_H(j,u)$ is the hamming distance between the bits representing S_j and S_u .

3.14 Proposed System model:

The simulated system model that used in this study can be illustrated as in figure (3.7) below:

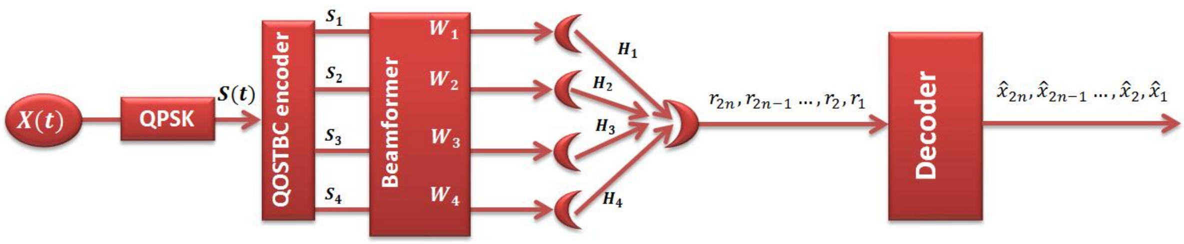

Figure (3.7): proposed system model

We assumed a flat fading channel which consists of L spatially separated paths, in our proposed system model L refers to the number of transmitter which is equals 4. The fading coefficients is given as h_l , and the Direction-Of-Arrivals (DOA) at the l^{th} path is given as θ_l then the linear combination of our received signal can be modelled as:

$$Y(t) = S(t)H^T W_t^H a(\theta_l) + n(t) \qquad (3.29)$$

$$Y(t) = \begin{bmatrix} r_1 \\ r_2 \\ r_3 \\ r_4 \end{bmatrix} = \begin{bmatrix} s_1 & s_2 & s_3 & s_4 \\ -s_2^* & s_1^* & -s_4^* & s_3^* \\ -s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & -s_3 & -s_2 & s \end{bmatrix} \begin{bmatrix} h_1.w_1.a(\theta_1) \\ h_2.w_2.a(\theta_2) \\ h_3.w_3.a(\theta_3) \\ h_4.w_4.a(\theta_4) \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \\ n_3 \\ n_4 \end{bmatrix} \qquad (3.30)$$

Where r_i is the received signal at the i^{th} time slot, and $a(\theta_l)$ is the downlink steering vector at θ_l, and n_i is the additive Gaussian noise at the i^{th} time slot.

The beamforming weight vector is set as w_i at the i^{th} beamformer. The signals corresponding to each row are transmitted at the same time slot [41]. Then the beamforming weight vector w_i is given by using the steering vector $k \times a(\theta_i)$, where k is a constant which assumed as:

$k \times a(\theta_i)$, where k is a constant which assumed as:

$$k = w_i^H \cdot a(\theta_i) = w_i^T \cdot a^*(\theta_i), (i = 1, \dots, 4) \quad (3.31)$$

Here $(\cdot)^T$ denotes matrix transpose operation, and $(\cdot)^H$ denotes the Hermitian matrix transpose or the matrix conjugate transpose operation.

3.15 Decoding:

The decoding is established using the same decoding method, which is ML but with low decoding complexity, as introduced and inspired from a new approach for QOSTBC as follow [42]:

First: let's mention the proposed coding system presumptions, the QOSTBC used is as:

$$A = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix},\ B = \begin{bmatrix} x_3 & x_4 \\ -x_4^* & x_3^* \end{bmatrix} \quad (3.32)$$

And the coding is assumed to be matrix as:

$$\mathrm{C} = \begin{bmatrix} \mathrm{A} & \mathrm{B} \\ -\mathrm{B}^* & \mathrm{A}^* \end{bmatrix} = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \\ -x_2^* & x_1^* & -x_4^* & x_3^* \\ -x_3^* & -x_4^* & x_1^* & x_2^* \\ x_4 & -x_3 & -x_2 & x_1 \end{bmatrix} \quad (3.33)$$

Second: The Equivalent Virtual Channel Matrix (EVCM) for this coding matrix above is denoted as H_v and can be given as:

$$H_v = \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2^* & -h_1^* & h_4^* & -h_3^* \\ h_3^* & h_4^* & -h_1^* & -h_2^* \\ h_4 & -h_3 & -h_2 & h_1 \end{bmatrix} \quad (3.34)$$

H_v can be described as a highly structured, equivalent, virtual (4x4) MIMO channel matrix that replaces the (4x1) received channel vector Y.

Third: establish the preceding method by multiplying the received vector Y with H_v^H as:

$$\widehat{\boldsymbol{X}} = H_v^H Y = H_v^H \cdot H_v \mathrm{C} + H_v^H n = D_4 \mathrm{C} + H_v^H n \quad (3.35)$$

Where $D_4 = H_v^H \cdot H_v$, is the detection matrix used to decode the received signal, H_v^H is the Hermitian of H_v , C is the QOSTBC coding matrix, and n is the noise vector of the AWGN channel.

For the OSTBC scheme the detection matrix is always diagonal this enables the use of simple linear decoding, but in the QOSTBC scheme this cannot be done due to the non-orthogonal detection matrix, as shown in below:

$$D_4 = \mathrm{C}^H \mathrm{C} = \begin{bmatrix} a & 0 & 0 & b \\ 0 & a & -b & 0 \\ 0 & -b & a & 0 \\ b & 0 & 0 & a \end{bmatrix} \quad (3.36)$$

Where the diagonal elements a represent the channel gains and b represent the interference from the neighbouring signals, for four transmit antennas. Then (a) and (b) can be estimated as in (3.10) and (3.11) by:

$$a = |h_1|^2 + |h_2|^2 + |h_3|^2 + |h_4|^2 \quad (3.37)$$

$$b = h_1 h_4^* + h_4 h_1^* - h_2 h_3^* - h_3 h_2^* \quad (3.38)$$

Then the estimated recovered symbols are can be returned using the MRC with the equation:

$$\hat{\boldsymbol{X}} = (H_v^H \cdot H_v)^{-1} \cdot H_v^H Y = (H_v^H \cdot H_v)^{-1} H_v^H \cdot H_v \cdot \mathrm{C} + (H_v^H \cdot H_v)^{-1} H_v^H \cdot H_v \cdot n$$

Where $(\cdot)^{-1}$ denotes the reverse of the dot product result matrix, then the equation above is simplified with the following procedures to maintain the transmitted symbols as $\hat{X}_1, \hat{X}_2, \hat{X}_3, and\ \hat{X}_4$. A

$$A = a + b = \left[(h_1 + h_4) \times \overline{(h_1 + h_4)}\right] + \left[(h_2 - h_3) \times \overline{(h_2 - h_3)}\right] \quad (3.39)$$

$$B = a - b = \left[(h_1 - h_{\mathrm{AAA4}}) \times \overline{(h_1 - h_4)}\right] + \left[(h_2 + h_3) \times \overline{(h_2 + h_3)}\right] \quad (3.40)$$

Here will calculate a preceding values that are dependent for estimation the final results:

$$C_1 = \left[R_1\overline{(h_1 + h_4)} + \overline{R_2}(h_2 - h_3) + \overline{R_3}(h_3 - h_2) + R_4\overline{(h_1 + h_4)}\right]/A \quad (3.41)$$

$$C_4 = \left[R_1\overline{(h_1 - h_4)} + \overline{R_2}(h_2 + h_3) + \overline{R_3}(h_3 + h_2) + R_4\overline{(h_1 - h_4)}\right]/B \quad (3.42)$$

$$C_2 = \left[R_1\overline{(h_2 + h_3)} + \overline{R_2}(h_4 - h_1) + \overline{R_3}(h_4 - h_1) - R_4\overline{(h_2 + h_3)}\right]/B \quad (3.43)$$

$$C_3 = \left[R_1\overline{(h_2 - h_3)} - \overline{R_2}(h_4 + h_1) + \overline{R_3}(h_1 + h_4) + R_4\overline{(h_2 - h_3)}\right]/A \quad (3.44)$$

And finally the recovered symbols are maintained as that, X_1, X_4 from C_1, C_4 , and X_2, X_3 from C_2, C_3 as given by:

$$\widehat{X}_1 = (C_1 + C_4)/2 \quad (3.45)$$

$$\hat{X}_4 = C_1 - \hat{X}_1 \quad (3.46)$$

$$\hat{X}_2 = (C_2 + C_3)/2 \quad (3.47)$$

$$\hat{X}_3 = C_2 - \hat{X}_2 \quad (3.48)$$

These equations are exactly the simplified ML algorithm for the proposed system model decoding method, which has a generalized structure described in the ML cost equations (3.17) and (3.18) described before.