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Abstract : The basic idea of COFDM is to split the modulation samples of incoming data stream onto a large number of carriers instead of modulating a unique carrier. Therefore, COFDM is an effective technique for combating multi-path fading and for high-bit-rate transmission over wireless channel. In a single carrier system a frequency Selective fading can cause the entire transmission link to fail, but in an COFDM multi carrier system, only a small percentage of the sub-carriers will be corrupted. Frequency and time interleaving in conjunction with forward error correction coding can then be used to correct for erroneous sub-carriers. The background information with the aim to provide an intuitive explanation of our research motivation.

C-OFDM is the modulation scheme of choice , as enshrined in International standard for all forms of digital broadcasting both audio and video and including satellite, terrestrial, and cable. In the existing standard the "coding" referred to consists of an inner convolutional code concatenated with an outer R-S code; here in this thesis, we replace the inner code with the coding like space time trellis code for analysis

Key Words : COFDM, STC, CSI, Interleaver

Introduction:

COFDM:

COFDM is a form of modulation which is particularly well-suited to the needs of terrestrial broadcasting channel. COFDM can cope with high levels of multi-path propagation , with a wide spread of delays between the received signals. It also cope with co-channel narrowband interference. COFDM has therefore been chosen for two recent new standards for broadcasting-DAB and DVB-T, both of which have been optimized for their respective applications and have options to suit particular needs. The special performance of COFDM in respect of multi-path and interference is only achieved by a careful choice of parameters and with attention to detail in the way in which the forward error-correction coding is applied. COFDM involves modulating the data onto a large number of carriers using the FDM technique. The key features which make it work, in a manner that is so well suited to terrestrial channels, include: Orthogonality (the "O" of COFDM),the addition of a Guard interval, the use of Error coding(the "C" of COFDM), Interleaving and Channel-State-Information.

The Importance of Orthogonality: The "Orthogonal" part of the COFDM name indicates that there is a precise mathematical relationship between the frequencies of the carriers in the system. In a normal FDM system, the many carriers are spaced apart in such a way that the signals can be received by using conventional filters and demodulators. In such receivers guard bands have to be introduced between different carriers and these guard bands in frequency domain results in a lowering of the spectrum efficiency. It is possible, however, to arrange the carriers in a COFDM signal so that the sidebands `of the individual carriers overlap and the signals can still be received without adjacent carrier interference . In order to do this, the carriers must be mathematically orthogonal.

Addition of Guard Interval: In practice, our carriers are modulated by complex numbers which change from symbol to symbol. If the integration period spans two symbols, not only will there same carrier ISI, but in addition there will be ICI as well. This happens because the beat tones from other carriers may no longer integrate to zero if they change in phase and/or amplitude during the period. we avoid this by adding a guard interval, which ensures that all the information integrated comes from the symbol and appears constant during it. As long as the delay of any path with respect to the main path is less than the guard interval, all the signal components within the integration period come from the same symbol and the orthogonality criterion is satisfied. ICI and ISI will only occur when the relative delay exceeds the guard interval.

Need of Error Coding: In fact, we would expect to use forward error-correction coding in almost any practical digital communication system, in order to be able to deliver an acceptable BER at a reasonably low SNR. At a high SNR it might not be necessary-and this is also true for uncoded OFDM, but only when the channel is relatively flat. Uncoded OFDM does not perform well in a selective channel. Again, for any reasonable number of carriers, CW interference that is affecting one carrier is less of a problem than a 0dB echo. However, just adding hard-decision-based coding to this Uncoded system is not enough,either-it would take a remarkably powerful hard-decision code to cope with an SER of 1 in 4! By using an error-correcting code which adds extra bits at the Transmitter it is possible to correct many or all of the bits received incorrectly.

Channel coding and Interleaving: OFDM is invariably used in conjunction with channel coding (forward error correction) and almost always uses frequency and/or time interleaving. Frequency interleaving increases resistance to frequency-selective channel conditions whereas time interleaving is of little benefit in slowly fading channels. The reason why interleaving is used on OFDM is to attempt to spread the errors out in the bit-stream that is presented to the error correction decoder, because when such decoders are presented with a high concentration of errors the decoder is unable to correct all the bit errors, and a burst of uncorrected errors occurs. A common type of error correction coding used with COFDM-based systems is Convolutional coding, which is often concatenated with Reed-Solomon coding. Convolutional coding is used as inner code and reed-solomon coding is used as outer code- usually with additional interleaving in between the two layers of coding. The reason why this combination of error correction coding is used is that the Viterbi decoder used for convolutional decoding produces short errors bursts when there is a high concentration of errors, and Reed-Solomon codes are inherently well-suited to correcting burst of errors.

Soft Decision Decoding and CSI: A hard-decision receiver would operate according to the rule that negative signals should be decoded as “0” and positive ones as “1”, with 0V being the decision boundary. If the instantaneous amplitude of the noise were never exceed ±1, then this simple receiver would make no mistakes. But noise may occasionally have a large amplitude, although with lower probability than for smaller values. Thus if say,+0.5V is received, it most probably means that a “1” was transmitted, but there is a smaller yet still finite probability that actually “0” was sent. This view of a degree of confidence is exploited in soft-decision Viterbi decoders. These maintain a history of many possible transmitted sequences, building up a view of their relative likelihoods and finally selecting the values “0” or “1” for each bit, according to which has the maximum likelihood. For convenience, a Viterbi decoder adds logarithmic likelihoods to accumulate the likelihood of each sequence. The appropriate log-likelihood measure or metric of the certainty of each decision is indeed simply proportional to the distance from the decision boundary. The slope of this linear relationship itself also depends directly on the SNR.

When data are modulated onto the multiple COFDM carriers, the metrics become slightly more complicated as the various carriers will have different SNRs. For examples, a carrier which falls into a notch in the frequency response will comprise mostly noise; one in a peak will suffer much less. Thus, in addition to the symbol-by-symbol variations, there is another factor to take account of in the soft-decisions: data conveyed by carriers having a high SNR are a priori more reliable than those conveyed by carriers having low SNR. This extra a priori information is usually known as Channel-state-Information (CSI)

Space Time Codes: In space-time coding, multiple antennas are used at the transmitter. Coding of symbols across space and time can be employed to yield coding gain and diversity gain. Coding gain is defined as the reduction in signal to noise (SNR) for the same FER that can be realized through the use of a code . Diversity gain performance improvement that can be achieved from a system by using diversity. Figure 1.2 illustrates typical coding gain and diversity gain. The x-axis of the plot is frame error rate (FER) and y-axis is SNR. FER is a commonly used the measure of the system performance. It is the ratio of the number of erroneous frames of

data at the receiver output to the total number of total transmitted frames of data. Space-Time Codes (STC) were introduced independently by Tarokh and Alamouti as a novel means of providing transmit diversity for multiple-antenna fading channels. There are two different types of STC. One is as space-time trellis codes (STTCs) and other is space-time block codes (STBCs) In addition decoding a STC is complicated. To reduce the decoding complexity, Alamouti discovered a remarkable scheme for transmitting using two transmit antennas, which is appealing in terms of both simplicity and performance.

Tarokh et al. used this scheme for an arbitrary number of transmitter antennas, leading to the concept of space-time block codes (STBC). STBC s have a fast decoding algorithm. STBC s can be classified into real orthogonal designs and complex orthogonal designs. The former deals with real constellations such as PAM, while the latter deals with complex constellations such as PSK and QAM. Real orthogonal designs have been well developed. In, Tarokh et al. proposed systematic constructions of real orthogonal designs for any number of transmit antennas with full rate. However, complex orthogonal designs are not well understood.

There exist several different types of space-time block codes obtained from complex orthogonal designs .One of the key features the STBC s obtained from orthogonal designs is that by performing linear processing at the receiver data symbols can be recovered. This feature is attractive for mobile and portable communication systems. The diversity gain obtained from a SITC is equal to the diversity gain from a STBC for the same numbers of transmit and receive antennas. However S'ITC can also provide coding gain, which a STBC cannot provide. However this additional coding gain is obtained at the cost of increased decoding complexity at the receiver because a Viterbi or trellis based decoder has to be employed. Note that the complexity of the decoder increases with the number of states in the trellis and the number of transmit antennas

SYSTEM MODEL

Time Domain OFDM System Model:

In order to understand the mathematical principles of OFDM systems, let's first take an overview of the time domain uncoded OFDM system model which is shown in Fig given below..

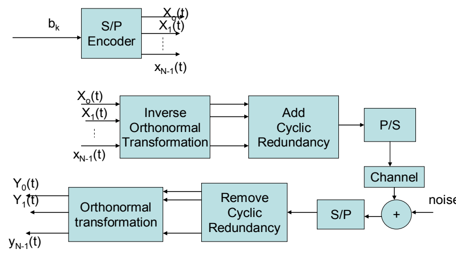

Fig.(1) OFDM System Model

Let b_k represent the binary data sequence to be transmitted over the channel. This data is divided into non-overlapping blocks of $n=N.\log_2 M$

bits. The n bits of data are partitioned into N-groups, with each $\log_2 M$ bits maps into a complex symbol of constellation size M. Symbol $x_k^{(t)}$ is the signal transmitted over the k^{th} sub-carrier during the t^{th} OFDM frame. At the transmitter, IDFT is performed as a method of modulation which results in samples $X_k^{(t)}$ given by ,

$$X_k^{(t)} = 1/\sqrt{N}.\sum_{i=0}^{N-1} x_i^{(t)} .\exp(j2\prod ki/N \qquad 0\leq i\leq N-1 \text{-----------}(1)$$

Assume the channel is frequency-selective and hence the implementation of a cyclic redundancy of sufficient length to the N-point OFDM frame is an effective method to counter ISI. The cyclic prefix causes the sequence $\{X_k^{(t)}\}$ to appear periodic to the channel and clears the channel memory at the end of each OFDM frame. This action makes successive transmission independent.

The output from the channel with additive noise $N_k^{(t)}$ may be written as ,

$$Y_k^{(t)} = C_k^{(t)} + X_k^{(t)} + N_k^{(t)} \text{ ……………..}(2)$$

Where, $C_k^{(t)}$ is the discretized fading channel co-efficient.

At the receiver the cyclic prefix is discarded to obtain a frame of N symbols $Y_k^{(t)}$. Taking N-point DFT we've output samples given by,

$$y_k^{(t)} = 1/\sqrt{N}.\sum^{N-1} Y_i^{(t)} .\exp(-j2\prod ki/N),$$

i=0 $\quad 0 \le i \le N-1$ …….(3)

Equivalent Frequency-Domain C-OFDM System Model:

Although the time domain model provided in the previous section is conceptually straightforward, it is much more insightful to analyze the OFDM system in the frequency domain. Hence let's consider the frequency model for a C-OFDM system as shown

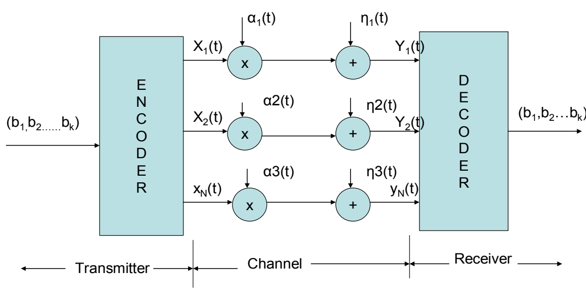

Fig.(2) C-OFDM System Model

A block of k information bits denoted as b= $(b_1,b_2\ldots b_k)$ is encoded into a codeword X $(x_1,x_2\ldots x_n)$ of length n , where each symbol x_i is an element from a complex alphabet X . There are a total of m-codewords in the codebook and the code rate is defined to be $R=(\log_2 m)/n$. Note that here we combine encoder, modulator and interleaver together to form one super encoder E. The encoded block X $(x_1,x_2\ldots x_n)$ is segmented into 'l' frames each of length N (lN=n). The individual frame is transmitted by N dependent parallel sub-channels, each representing a different OFDM sub-carrier. According to the tapped-delay-line model the fading co-efficient $\alpha_i^{(t)}$ of the i^{th} OFDM frame are related to the fading envelopes $C_i^{(t)}$ through

$$C^{(t)} = [C_1^{(t)}, \ldots C_l^{(t)}, 0, \ldots\ldots 0]^T \in C^{N+1} \ldots (4)$$

$$\alpha^{(t)} = [\alpha_1^{(t)}, \ldots \alpha_N^{(t)}]^T \in C^{N+1} \quad \ldots\ldots\ldots (5)$$

$$\alpha^{(t)} = W_{NxN} . C^{(t)} \quad \ldots\ldots..(6)$$

where Fouirer transformation W_{NxN} is given by,

$$W_{NxN} = \begin{bmatrix} W^0 & W^0 \ldots\ldots\ldots & W^0 \\ W^0 & W^1 \ldots\ldots\ldots & W^{N-1} \\ W^0 & W^{N-1} \ldots\ldots\ldots & W^{(N-1)(N-1)} \end{bmatrix}$$

$$W = e^{-j2\pi/N} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(7)$$

Each component $C_i^{(t)}$ of the fading envelope is assumed to be independent from tap-to-tap. A rayleigh fading distribution is considered in this paper with probability density function of $[C_i^{(t)}]$ is given by,

$$F_{C_i^{(t}}(x) = x/\sigma^2 \exp(-x^2/2\sigma^2), \quad x \ge 0 \ldots\ldots\ldots\ldots\ldots\ldots(8)$$

Where, $E[\ |C_i^{(t)}|^2] = 2\sigma^2$ is the average power of the fading envelope. It is further assumed that each tap has same average power.

The receiver o/p vector $y^{(t)}$ is given by,

$$y^{(t)} = [y_1^{(t)}, y_2^{(t)}, \ldots\ldots y_N^{(t)}]^T \ldots\ldots\ldots\ldots\ldots\ldots(9)$$

$$y_i^{(t)} = \alpha_i^{(t)} . x_i^{(t)} + n_i^{(t)} \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(10)$$

where additive complex Gaussian noise $n_i^{(t)}$ is white with varience N_0 .

The receiver is assumed to have perfect knowledge of channel state information and makes the decision based on the observations on $y^{(t)}$ and CSI, C.

System Model for STC:

We consider a K-subcarriers OFDM system with N transmit antennas and M receive antennas. The data to be transmitted enters the space-time trellis encoder that will produce n streams of modulated constellation symbols. At any given time t, the outputs of space-time trellis encoder, $x_t^1, x_t^2, \ldots.., x_t^n$ can be expressed as

$(x_t^1, x_t^2, \ldots.., x_t^n) =$ aG (mod M) -------------(11)

For the 4-PSK ($M = 4$), 4 state STTC's, the input bits to the space-time trellis encoder at time t are (a_t^1, a_t^2) and $a = (a_t^1, a_t^2, a_{t-1}^1, a_{t-2}^2)$. The generator matrix G is a $4 \times N$ matrix for the 4-PSK, 4 state STTC's where each element of G, $g_{k,l}$,is $g_{k,i} = 0, 1, 2, 3$ for the 4-PSK. For SFTC's, the output codeword is modulated onto one subcarrier of the OFDM word of it's corresponding transmit antenna. The symbol at the k-th ($k = 0, 1, \ldots\ldots, K-1$) subcarrirer transmitted through the i-th

(i = 1, 2, ………, N) antenna is denoted by $c_i[k]$. We append a cyclic prefix to each OFDM frame to avoid the possible ISI due to the delay spread. At the receiver, after matched filtering, sampling and OFDM demodulation, the received signal at the j-th receive antenna (j = 1, 2, ……., M) and the k-th subcarrier can be expressed as

$$r_j[k] = \sum_{i=1}^{N} H_{i,j}[k]\, c_i[k] + \eta_j[k] \text{ ---------------(12)}$$

where $H_{i,j}[k]$ denotes the normalized channel frequency response

from the transmit antenna i to the receive antenna j at the k-th subcarrier. The noise $\eta_j[k]$ at the j-th receive antenna and the k-th subcarrier is modeled as independent samples of a complex Gaussian random variable with mean zero and variance $N0/2$ per dimension. The frequency selective fading channel is modeled as a tapped-delay-line. The time-domain channel impulse response between the transmit antenna i and the receive antenna j can be expressed as

$$h_{i,j}(t, \tau) = \sum_{l=1}^{L} \alpha_{i,j}(t, l)\delta(\tau - n_l T_s) \text{ ------------------(13)}$$

where L is the number of non-zero taps, $\alpha_{i,j}(t, l)$ is the complex

coefficient of the l-th tap with an integer delay n_l, and T_s is the sampling interval of the OFDM system. Moreover, $\alpha_{i,j}(t, l)$ are modeled by the wide-sense stationary (WSS) narrowband complex Gaussian processes, which are independent for different path's with $E[|\alpha_{i,j}(t, l)|^2] = \sigma_l^2$. We normalize the channel power by letting $\Sigma\ \sigma_l^2 = 1$. Since we consider a -quasi-static fading channel, i.e., the fading coefficients are constant during an OFDM frame, the time index can be dropped for brevity. With proper cyclic prefix, perfect sampling time and tolerable leakage, the channel frequency response at the k-th sub-carrier between the transmit antenna i and the receive antenna j can be expressed as

$$H_{i,j}[k] = \sum_{l=1}^{L} h_{i,j}[l] \exp(-j2\pi k n_l/k) \text{ --------- (14)}$$

We assume that the receiver can obtain the perfect channel state information. We also assume that a codeword defined as $\mathbf{c} = (\mathbf{c}[0], \ldots, \mathbf{c}[k], \ldots, \mathbf{c}[K-1]) = (c_1[0]c_2[0] \ldots c_N[0], \ldots, c_1[k]c_2[k] \ldots c_N[k], \ldots, c_1[K-1]c_2[K-1] \ldots c_N[K-1])$ was transmitted and erroneously decoded as $\mathbf{e} = (\mathbf{e}[0], \ldots, \mathbf{e}[k], \ldots, \mathbf{e}[K-1]) = (e_1[0]e_2[0] \ldots e_N[0], \ldots, e_1[k]e_2[k] \cdot \ldots e_N[k], \ldots, e_1[K-1]\ e_2[K-1] \ldots e_N[K-1])$. The pairwise error probability (PWEP) of deciding in favor of $\mathbf{e}$ using maximum likelihood decoder, conditioned on $\mathbf{H}_{i,j,k} = (H_{i,j}[0]H_{i,j}[1] \ldots H_{i,j}[K-1])$, is upper bounded as $\Pr(\mathbf{c} \to \mathbf{e}|H_{i,j,k}) \le \exp(-Es/4No * d2(c,e))$- -------(15)

where Es is the energy per symbol at each transmit antenna and $N0$ is the noise power spectral density.

$d^2(\mathbf{c}, \mathbf{e})$ is given by

$$d^2(c,e) = \sum_{j=1}^{M} \sum_{k=0}^{K-1} \left| \sum_{i=1}^{N} H_{i,j}[k]\, q_i[k] \right|^2$$

with $q_i[k] = c_i[k] - e_i[k]$, $\mathbf{q}[k] = [q1[k], \ldots, qN[k]]T$,

We assume that the fading gains follow a Rayleigh distribution.

By averaging (5) with respect to the Rayleigh distribution, the

PWEP is upper bounded as

$$\Pr(\mathbf{c} \to \mathbf{e}) \le \prod_{K=0}^{K-1} (1 + |c[k]-e[k]|^2 E_s/4No)^{-M} \text{-------(16)}$$

Here, we consider the bound when the diversity gain is moderate, or equivalently, the required SNR is moderate. Assuming that the required SNR is moderate, it follows from (6) that

$$\Pr(\mathbf{c} \to \mathbf{e}) \le \prod_{k=0}^{K-1} (1 + |\mathbf{c}[k] - \mathbf{e}[k]|^2 Es/4N_o)^{M} \text{---(17)}$$

$$= \left(1 + \sum_{k=0}^{K-1} |\mathbf{c}[k] - \mathbf{e}[k]|^2 Es/4N_o + o(E_s/4N_o)\right)^{-M}$$

$$\approx \left(\sum_{k=0}^{K-1} |\mathbf{c}[k] - \mathbf{e}[k]|^2 Es/4N_o\right)^{-M} \text{---------(18)}$$

where $o(E_s/4N_o$) denotes the summation of all the terms that include higher order quantities of ($Es/4N0$).

!&**The Trace Criterion**: To achieve the largest possible coding gain, the minimum value of the trace of the codeword distance matrix on STTC's

$$tr_{min} = \min \sum_{k=0}^{K-1} |c[K] - e[K]|^2 \text{--------}(19)$$

between any two distinct codewords ***c*** and ***e*** must be maximized.

Significance of the Research:

As mentioned earlier, STBC s provide diversity gain but no coding gain. On the other hand SITC s have both diversity and coding gain, but the complexity of designing a good code is the main drawback of STTC. There has been rapid progress in this field, targeted at finding better codes with full diversity and with greater coding gain than those provided earlier. Improved S'ITCs that were found through exhaustive computer search over a feed-forward convolution code (FFC) generator. Ionescu et al. found improved 8 and 16-state S'ITC s for 4-PSK for the case of two transmitters in flat Rayleigh fading via a modified determinant criterion. Similarly Chen et al. derived more accurate code design criteria by using a tighter bound for the ***Q(.)*** function in the pairwise error probability (PEP) approximation in .

This yielded new S'ITCs with better performance than the original codes proposed by Tarokh et al. A more structured method of code construction that ensures full diversity is provided , along with a number of new code designs, such as codes that yield the best distance spectrum properties among all codes with a given coding gain [%]. Boleskei et al. considered the effect of receive and transmit correlation in multiple-input multiple-output (MIMO) systems on error performance of STCs. They showed that the resulting maximum diversity order was given by the ranks of the receive and transmit correlation matrices. Further work has been undertaken to study the performance of STBCs and STTCs and to develop robust codes for correlated fading channels . The work in this thesis concentrates on the performance of the codes over Rayleigh fading channel. In this thesis we first present the performance of OFDM system over fading channel and then we compare these results with the performance in C-OFDM system.

Here we are adopting 4-PSK space time trellis codes for performance analysis. Then a comparison will be made on bit error rate and BER v/s SNR between OFDM and C-OFDM systems.

Here, we present a rationale for STTC performance over Rayleigh fading channel.

System Model of STTC Based Wireless System

A typical S'ITC based wireless system has an encoder, pulse shaper, modulator and multiple transmit antennas at the transmitter, and the receiver has one or more receive antennas, demodulator, channel estimator and S'ITC decoder. We consider a mobile communication system with n_T transmit antennas and n_R receive antennas as shown in Fig. (3) and (4). The space-time trellis encoder encodes the data ***s(t)*** coming from the information source and the encoded data is divided into n_T streams of data $c_t^1, c_t^2, \ldots c_t^{n_T}$. Each of these streams of data passes through a pulse shaper before being modulated. The output of modulator i at time slot t is the signal c_t^i , which transmitted through is transmit antenna i. Here $1 \le i \le nT$. The transmitted symbols have energy E_s . We assume that the nT signals are transmitted simultaneously from the antennas. The signals have transmission period *T.* In the receiver, each antenna receives a superposition of nT transmitted signals corrupted by noise and multipath fading. Let the complex channel coefficient between transmit antenna i and receive antenna j have a value of ***hi,j(t)*** a t time ***t*** ,w here $1 \le j \le n_R$

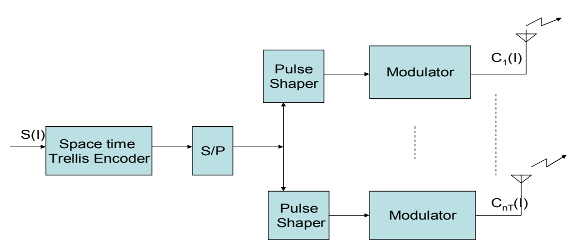

Fig.(3) A block diagram of a STTC based system.

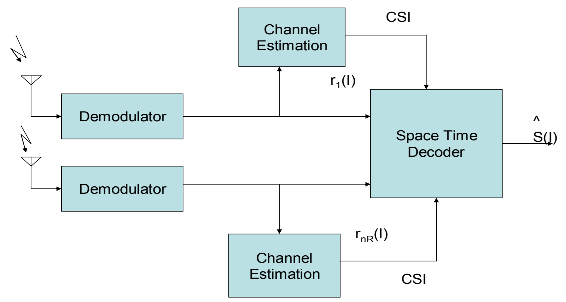

Fig.(4) A block diagram of receiver of a STTC based system.

The received signal at antenna j, j = 1,2, ..., n_T [I] is then

$$r_t^j = \sqrt{E_s} \sum_{i=1}^{n} h_{i,j}(t)\, c_t^j(t) + \eta_t^j \qquad \text{----------------(20)}$$

where η_t^j is additive white Gaussian noise (AWGN) at receive antenna ***j*** **,** which has zero mean and power spectral density ***No*** and $h_{i,j}(t)$ is the channel coefficient between transmit and receive antennas .

Code Construction of 4-state 4-PSK STTC

A signal constellation diagram for 4-PSK is shown in Fig.(5). With PSK information is contained in the signal phase. For 4-PSK, the phase takes one of four equally spaced values, such as 0, $2\pi / 4$, $4\pi/4$ and $6\pi / 4$. These are typically represented by a Gray code , as shown on the right side

of Figure. These signal points are also labeled as 0,1,2 and 3. We can also express these in complex notation. The encoder structure of a 4-state 4-PSK STTC is shown in Fig.(6), with bits input to the upper and lower branches. The memory order of the upper and lower branches are ***u*,** and ***v*,** ,

respectively. These are basically shift registers. The main purpose of the shift registers in the encoder is to store the previous transmitted bits. The length of the shift register is the memory of the encoder. The branch coefficients are arranged alternatively in the generator matrix, with a_i representing the most significant bit (MSB). The input bit streams I_t^1 and I_t^2 are fed into the branches of the encoder with I_t^1 being the MSB. The output of the encoder is

$$x_t^k = \left(\sum_{p=0}^{v1} I_{t-p}^1 a_p^k + \sum_{q=0}^{v2} I_{t-q}^2 b_q^k \right) \bmod 4 \qquad k=1,2\text{------(21)}$$

where $v1 + v2 = v$ and the number of states is 2^v . vi is calculated as

$$v_i = [\,(v+i-1)/2\,], \quad i=1,2$$

Here [x] denotes the largest integer smaller than or equal to x. For each branch, the output is the sum of the current input scaled by a coefficient and the previous input scaled by another coefficient. The two streams of input bits are passed through their respective shift register branches and multiplied by the coefficient pairs (a_p^1, a_p^2) and (b_q^1, b_q^2) . Here $a_p^k\, b_q^k \in \{0,1,2,3\}$, k=1,2 p=0,1,2…….,v1 and q=0,1,2,…..v2

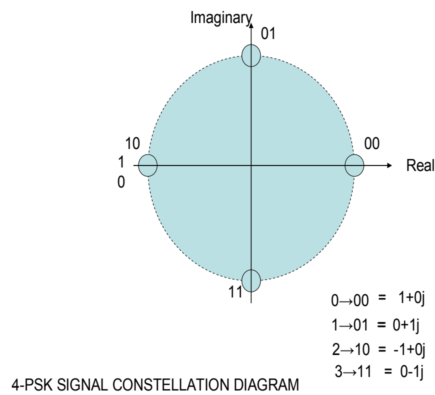

Fig.(5) 4-PSK signal constellation diagram

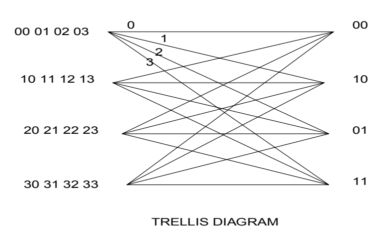

Fig.(6) 4 PSK 4-state STTC Trellis diagram

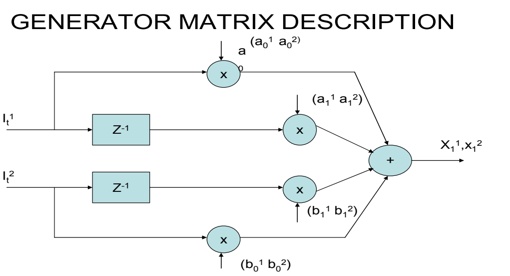

Fig.(7) Encoder Structure

Then x_t^1 and x_t^2 are transmitted simultaneous through the first and second transmit antennas, respectively.

Design Criteria for STTC over Rayleigh Fading

Euclidean distance Criterion:

When the diversity gain is large (with two or more receive antennas), proposes the design criteria, namely the Euclidean Distance Criteria (EDC). Usually the Rank and Determinant criteria (RDC) applies to the systems with a single receive antenna and a small number of transmit antennas. This shows that with diversity gain $rn_R \geq 4$, shows that the error probability is upperbounded by

$$P_e(c \rightarrow e) \leq 1/4 .\exp(-n_R.E_s/4N_o \sum_{i=1}^{n_T} \sum_{j=1}^{l} | e_j^i - c_j^i |) \quad \text{when } rn_R \geq 4 \text{ --------------(22)}$$

which indicates that we should maximize the minimum squared Euclidean distance between any two different codewords .

STTC Decoder

The decoder is based on the Viterbi algorithm, so it uses the trellis structure of the code. Each time the decoder receives a pair of channel symbols it computes a metric to measure the "distance" between what is received and all of the possible channel symbol pairs that could have been transmitted. For hard decision Viterbi decoding the Hamming distance is used, and the Euclidean distance is used for soft decision Viterbi decoding.

The metric values computed for the paths between the states at the previous time instant and the states at the current time instant are called "branch metrics". We assume that the decoder has ideal channel state information (CSI) and thus knows the path gains $h_{i,j}$, (where i = 1,2,. . . , n_T and j = 1,2,. . . , n_R). If the signal is r_t^j at receive antenna j and time t , the branch metric for a transition labeled $x_t^1, x_t^2 \ldots x_t^{n_T}$ is given by :

$$\sum_{j=1}^{n_R} | r_t^j - \sum_{i=1}^{n_T} h_{i,j} q_i^t |^2 \quad \text{---------------(23)}$$

The Viterbi algorithm determines the path with the lowest accumulated metric.

SIMULATION RESULT

Simulation Parameters:

In our simulations we considered the IS-136 standard . In this system, performance is measured by the frame error rate (FER) for a frame consisting of 130 symbols. We also assumed ideal channel state information (CSI) is available at the receiver. We carried out the simulation by MATLAB. Random M-PSK symbols are set in frames as a group, which consists of 130 symbols each. The space-time encoder takes the frame as input and generates codeword pairs of each input symbol simultaneously for all the transmit antennas. Pulse shaping and matched filter are used. These complex signals are transmitted through the MIMO channel.

In our simulations, two transmit and two receive antennas are assumed. A two equal-power taps ($L = 2$) fading channel with the delay spread of 5 μs is adopted. It is also assumed that the receiver can obtain perfect channel state information and that the fading gains are constant during each OFDM frame and vary from one frame to another.

In this paper first we have plotted the BER v/s SNR plot of OFDM as reference. Next we have evaluated BER performance using STTC code. Simultaneously, we have plotted graph for channel capacity v/s snr. The performance of STTC is evaluated with and without interleaving. By comparing the OFDM and resulted STTC-OFDM we can conclude that coded OFDM is giving the better result in comparison to OFDM.

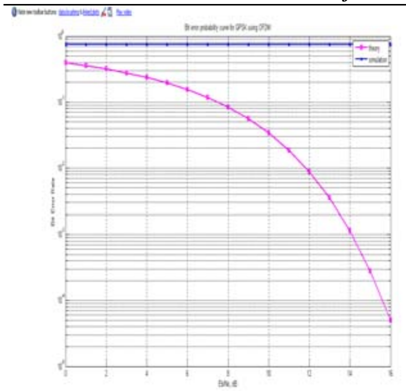

Fig.(8) Plot of BER v/s E_b/N_0 using OFDM

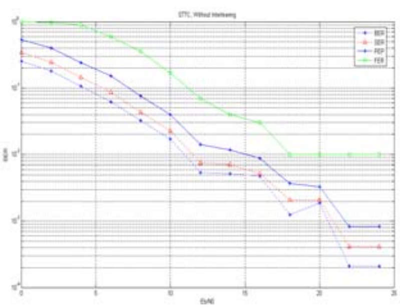

Fig.(9)Plot of BER,FER,SER and PEP v/s E_b/No (Logarithimic value) without interleaving.

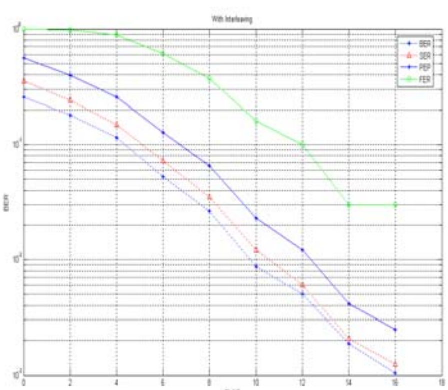

Fig.(10) Plot of BER,FER,SER and PEP v/s E_b/No (Logarithimic value) with interleaving.

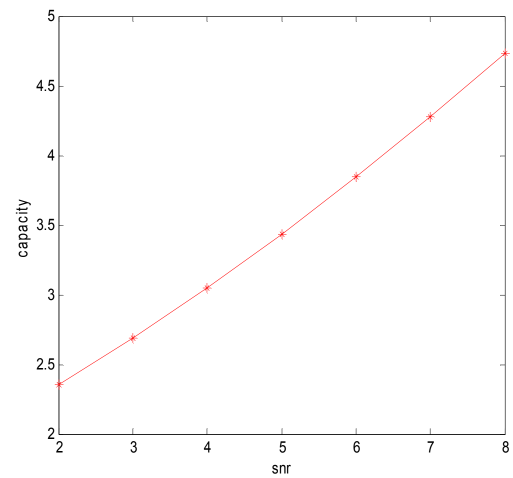

Fig.(11) Plot of Channel capacity v/s snr

Conclusion:

This thesis particularly based on the evaluation of BER performance in a cellular system using Coded OFDM technique. Here , we have taken general plot of BER v/s SNR using OFDM modulation scheme over Raleigh channel as reference. Then we applied coded OFDM technique, i.e. replacing convolution code with trellis code and finally evaluate the BER performance. Here, 4-state 4-PSK is taken as modulation scheme. We have adopted the Euclidean Distance Criterion for analysis. We have plotted the logarithmic value of BER ,FER, PEP and SER v/s E_b/N_o .At the same time we also plotted a graph which shows channel capacity increases proportionally with signal-to-noise ratio by using trellis coded OFDM. The performance of STTC is evaluated with and without interleaving. By comparing the OFDM and resulted STTC-OFDM we can conclude that coded OFDM is giving the better result in comparison to OFDM.But, so far Cellular system is concerned , it is a very important and vast area of research and providing scope for new and advanced techniques to cope with the problem associated with multi-path propagation.

My thesis can be expanded using 8-PSK modulation scheme. Also advanced coding techniques can be adopted. With Rank and Determinant Criterion (RDC) over rayleigh channel further research work can be made.

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