Enhanced Information Throughput in MIMO-OFDM based systems using Fractional Sampling and Iterative Signal Processing

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Abstract

Herein, we study the feasibility of enhanced information throughput capability of Multiple Input Multiple Output Orthogonal Frequency Division Multiplexing (MIMO-OFDM) based wireless communication systems in the context of increasing wireless capacity demand. The concept of Fractional Sampling (FS) is exploited for this purpose to take advantage of diversity created due to it. Furthermore, a novel channel adaptive iterative sequence detector is proposed using the FS technique. The performance analysis of the proposed receiver is carried out and a tighter performance bound is derived. It is found that the FS technique can improve Bit Error Rate (BER) performance of MIMO and OFDM based systems provided that the noise samples are uncorrelated up to a certain level of FS rate. Moreover, it is observed that the performance improvement is a non-linear function of the FS rate. Besides this, the simulation results show that the proposed iterative receiver can significantly enhance the information throughput of MIMO-OFDM based wireless communication systems in comparison to conventional non-iterative receivers.

Index Terms: MIMO, OFDM, Fractional Sampling, Iterative Detector, Information Throughput

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I. INTRODUCTION

Achieving high information throughput for the wireless systems operating in multi-path fading environments is a challenging task and it requires efficient transmission techniques to ensure reliable and efficient utilization of the available usable spectrum. Multiple antennas can be considered in both the transmitter and receiver to enhance the signal transmission reliability creating transmit and receive diversity [1], widely known as Multiple Input Multiple Output (MIMO) system. Furthermore, information throughput can be enhanced by spatial multiplexing of the information signals exploiting multiple antennas at the transmitter. The multi-carrier modulation scheme has been proposed in [2] to enhance the spectrum efficiency, known as Orthogonal Frequency Division Multiplexing (OFDM). The combination of OFDM with the MIMO system enhances the receiver Bit Error Rate (BER) performance due to the ease of equalization in the receiver as considered in the various literature [3–5]. For high data rate transmission, the multi-path characteristic of the environment causes wireless channels to be frequency selective. The OFDM technique can convert such a frequency selective MIMO fading channel into a set of parallel equivalent frequency flat MIMO fading channels, causing reduction in the receiver computational complexity. Subsequently, the combination of two powerful techniques, i.e., MIMO and OFDM, has become a promising accessing technology for broadband wireless access schemes such as 4G, IEEE 802.16, IEEE 802.11n, WiMax Mobile, WiMax Fixed, and 3GPP LTE [6–8]. However, exploring innovative techniques to enhance the information throughput of the next generation wireless systems in order to meet the throughput demand of the wireless users is still challenging and requires extensive research.

Fractional Sampling (FS) has been considered as an important research topic since many years for enhancing the signal detection reliability in the receiver [5, 9–11]. In the FS technique, the received signal is sampled with the rate higher than the Nyquist rate in order to achieve diversity within the received samples. In other words, a system with FS can be realized as a virtual multiple output system that has multiple independent channel paths with different fading effects. Experimental investigation of FS on a practical IEEE 802.11b WLAN system has been carried out in [10] and it is claimed that it has improved the performance up to 2 dB in non-line of sight conditions. However, the correlation of noise in parallel channels created due to FS limits the amount of diversity that can be achieved through the process of FS [12]. In the context of an FS OFDM receiver, a sampling point selection method based on the frequency response of the channel has been investigated in [11]. In [13], a low complexity receiver that exploits achievable diversity through FS has been proposed for a simple OFDM system. Furthermore, in [14, 15], an FS based Spectrum Sensing (SS) technique has been studied for a cognitive radio in the presence of noise correlation and it has been shown that the SS efficiency increases with the FS rate up to a certain limit. In the context of an FS based MIMO-OFDM system, the contribution in [5] uses the FS concept to enhance the channel capacity of a multiuser MIMO-OFDM system in the street-canyon channel. Subsequently, it has been shown that the capacity improves by 3.5 bps/Hz while using FS.

Iterative equalization has drawn the attention of many system designers and researchers in comparison to the traditional non-iterative detection techniques. Iterative detection can be used in several applications such as timing recovery, phase recovery, interference suppression, signal detection, equalization, decoding, multiuser detection [16, 17] etc. Recently, the iterative detection technique has received important attention in various aspects of the MIMO-OFDM system [4, 18–21].

In this paper, we propose an FS based MIMO-OFDM system and iterative signal processing at the receiver for enhancing the information throughput of the system. In the FS-OFDM model proposed in the literature [13], the received symbols are first converted to the frequency domain and combining is performed by using subcarrier by subcarrier combining approach. Each FS branch needs one OFDM demodulator, which increases the receiver complexity by the FS rate. In this paper, instead of combining in the frequency domain, time domain combining is considered which makes receiver circuitry simpler than the receiver proposed in [13]. Furthermore, the theoretical analysis of the proposed receiver structure for the FS based MIMO-OFDM system is presented and the performance of the system is evaluated in terms of probability of BER of the system. Moreover, a novel iterative detection scheme for MIMO-OFDM based systems based on FS is proposed. In many of the contributions found in the literature, the iterative detection process has been considered between a detector and a channel decoder. Furthermore, in conventional iterative detection process, an interleaver is used to generate signal diversity at the receiver as well as to prevent the occurrence of error bursting whereas in the proposed detection method, an interleaver is not needed as signal diversity is obtained with FS. According to author's knowledge, no work has been reported in the literature considering the iterative receiver using the proposed FS based method. The proposed detector uses an iterative detection process between the detectors, each detector being fed from different fractionally sampled version of the same sequence. This is the main contribution of this paper.

The rest of the paper is organized as follows: Section II provides the description of the proposed MIMO-OFDM system model along with its mathematical model. Section III presents the derivation of received SNR for MIMO-OFDM system with FS. Section IV describes the proposed iterative signal detection process. Section V provides the BER analysis for iterative detection in Additive White Gaussian Noise (AWGN) and Rayleigh fading channels. Section VI analyzes the performance of the proposed receiver with the help of numerical results. Section VII concludes this paper.

II. Proposed MIMO-OFDM system with FS

A. MIMO-OFDM Transmitter System Model

A MIMO-OFDM system with N_T number of transmit antennas and N_R number of receive antennas has been considered. The system model of the MIMO-OFDM transmitter is shown in Fig. 1. At the transmitter, the data stream is divided into different sub-streams, which are transmitted through different antennas. Denoting one block of the data sub-stream from the m-th transmit antenna before OFDM symbol mapping by $\mathbf{s}_m$, which is a frequency domain signal vector, can be written as [1]:

$$\mathbf{s}_m = \begin{bmatrix} s_m[0] & s_m[1] & \cdots & s_m[N-1] \end{bmatrix}^T, \tag{1}$$

where $m \in 1, 2, 3, ..., N_T$ and N is the length of the data block. After performing N-point Inverse Discrete Fourier Transform (IDFT), time domain signal vector, denoted by $\mathbf{v}_m$, can be written as [11]:

$$\mathbf{v}_m = \begin{bmatrix} v_m[0] & v_m[1] & \cdots & v_m[N-1] \end{bmatrix}^T = \mathbf{F}_N \mathbf{s}_m, \tag{2}$$

[1]For notational convenience, we drop the block index and we map each data block into one OFDM symbol.

where $\mathbf{F}_N$ is an $(N \times N)$ IDFT matrix with $(n+1, k+1)$-th entry being $e^{\frac{j2\pi nk}{N}}/\sqrt{N}$ and $n, k \in \{1, 2, ..., N-1\}$.

The Cyclic Prefix (CP) of length l_{cp} is added to a data sub-stream of length N bits to obtain an OFDM symbol. The CP length is expected to be greater than channel length to combat the effect of Intersymbol Interference (ISI). After adding CP, the corresponding OFDM symbol at the input of m-th transmit antenna can be denoted by $\mathbf{u}_m$ and can be written as:

$$\mathbf{u}_m = \begin{bmatrix} u_m[0] & u_m[1] & \cdots & u_m[N'-1] \end{bmatrix}^T, \tag{3}$$

where $N' = N + l_{cp}$. The transmitted continuous-time signal in baseband form is given by [11];

$$\mathbf{x}_m(t) = \sum_{n=0}^{N'-1} u_m[n]p(t - nT_s), \tag{4}$$

where $p(t)$ is the impulse response of pulse shaping filter and $1/T_s$ is the baud rate. As $\mathbf{x}_m(t)$ is passed through the channel, it is expected to be affected with channel response $c(t)$.

B. *MIMO-OFDM Receiver System Model*

In the receiver, firstly, the received signal is passed through a matched filter with response $p(-t)$. The received signal corresponding to the transmitted OFDM symbol at the input of r-th receive antenna can be written as [11]:

$$y_r(t) = \sum_{m=1}^{N_T} \sum_{n=0}^{N'-1} u_m[n]h(t - nT_s) + z_r(t), \tag{5}$$

where $z_r(t)$ is Gaussian noise at the output of matched filter at the r-th receive antenna branch, r ranges from 1 to N_R. The impulse response of the composite channel $h(t)$ is given by $h(t) = p(t) * c(t) * p(-t)$, where $(*)$ denotes the convolution operation. For a wireless multi-path channel, the impulse response $h(t)$ can be expressed in baseband form as:

$$h(t) = \sum_{i=0}^{N_h-1} \alpha_i p_1(t - \tau_i), \tag{6}$$

where α_i is the gain coefficient of i-th multi-path channel. $p_1(t) = \int p(t')p(t'+t)dt'$ is the deterministic correlation of $p(t)$. It is assumed that the channel has N_h complex gain coefficients and time invariant during each OFDM symbol. The delays τ_i are usually assumed to be uniformly distributed in $(0, \tau_{max})$, where τ_{max} is the maximum multipath spread of the channel.

The selection of sampling rate plays a critical role in order to maintain the white nature of the noise samples. If sampling rate of the received data is more than a certain threshold [2], then the noise samples obtained after the FS operation are expected to be correlated and the white noise becomes colored. In that case, a noise whitening filter is required after performing the FS [13]. In this analysis, it is assumed that sampling rate of the received data stream is lower than the threshold, hence the noise samples after performing FS can be assumed to be uncorrelated. Furthermore, we consider low FS rates for our analysis and at lower FS rates, the noise correlation effect is negligible as shown in our simulation results in Section VI.

After removing the CP, the output of the matched filter $y_r(t)$ is sampled at a rate of K/T_s, T_s being sampling interval and its components can be expressed as [13]:

$$b_r^k[n] = \sum_{m=1}^{N_T} \sum_{l=0}^{N-1} u_m[l] h_{r,m}^k[n-l] + z_r^k[n], \tag{7}$$

where

$$b_r^k[n] = b_r(nT_s + kT_s/K),$$

$$h_{r,m}^k[n] = h_{r,m}(nT_s + kT_s/K),$$

$$z_r^k[n] = z_r(nT_s + kT_s/K),$$

for $k = \{0, 1, ..., K-1\}$, where $b_r^k[n]$ for each value of k represents the circular convolution relationship between u_m and $h_{r,m}^k[n]$, $n = \{0, 1, ...N-1\}$. Equal Gain Combining (EGC) technique can be used to combine multi-path components arising from FS operation. In this work, time-domain combining has been considered, which makes receiver circuitry quite simpler than the model proposed in [13]. After performing DFT at the r-th receiver for each k, the following equation can be obtained:

$$\mathbf{p}_r[g] = \sum_{m=1}^{N_T} \mathbf{h}_{r,m}[g] s_m[g] + \mathbf{z}_r[g], \qquad g = 0, 1, ..., N-1 \tag{8}$$

where $\mathbf{p}_r[g] = [p_r^0[g], ..., p_r^{K-1}[g]]^T$ with $p_r^k[g] = N^{-1/2} \sum_{n=0}^{N-1} b_r[n] e^{-j2\pi gn/N}$, and similar equation can be written for $\mathbf{z}_r[g]$, $\mathbf{h}_{r,m}[g]$ is a $K \times 1$ column vector of the $K \times N$ matrix $\mathbf{H}_{r,m}$ and is given by;

$$\mathbf{h}_{r,m}[g] = \sum_{n=0}^{L-1} \mathbf{h}_{r,m}[n] e^{-j2\pi gn/N}, \tag{9}$$

[2]This threshold depends on the considered signal bandwidth.

where each pair of transmit and receive antennas is assumed to have L independent delay paths. Furthermore, the channel matrix $\mathbf{H}_{r,m}$ represents the (r,m)-th block of the block matrix $\mathbf{H}$ of dimension $N_R K \times N_T N$. It is assumed that the channel impulse response is known at the receiver. The equalized OFDM symbol vector at the receiver using Maximum Ratio Combining (MRC) can be written in the following generalized form:

$$\hat{\mathbf{s}} = \mathbf{s} + (\mathbf{H}^H\mathbf{H})^{-1}\mathbf{H}^H\mathbf{z}. \tag{10}$$

III. SNR OF THE MIMO-OFDM SYSTEM WITH FS

The output SNR for 2 transmit and 1 receive antenna (i.e., transmit diversity and no receive diversity) can be written in the following form [12]:

$$\lambda = \frac{1}{2}\sum_{l=1}^{L}[\lambda_{1,l} + \lambda_{2,l}], \tag{11}$$

where $\lambda_{q,l}$ is the average SNR of the l-th multipath channel from the q-th transmit antenna to the receive antenna. Factor $1/2$ is introduced due to division of power between two transmit antennas [22]. The same concept of calculating SNR can be extended to MIMO system having both transmit and receive diversity. The output SNR for N_T transmit and N_R receive antenna system can be written as [23]:

$$\lambda = \frac{1}{N_T}\sum_{i=1}^{N_T}\sum_{j=1}^{N_R}\sum_{l=1}^{L}\lambda_{i,j}^{l}, \tag{12}$$

where $\lambda_{i,j}^{l}$ is the average SNR from i-th transmit antenna to j-th receive antenna on l-th channel. Cyclic prefix is considered to be longer than the channel length. As long as CP is considered longer than the channel length, the frequency selective channel for OFDM case becomes frequency flat channel. After taking N point IDFT, the channel impulse response at the n-th frequency bin in terms of frequency domain can be expressed as [22]:

$$H_{i,r}(n) = \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1} h_{i,r}(k)e^{\frac{-j2\pi nk}{N}}, \tag{13}$$

for $0 \leq n \leq N-1$. The above equation can also be written in the following form:

$$H_{i,r}(n) = \frac{1}{\sqrt{N}}\mathbf{e}_n^H\mathbf{h}_{i,r}, \tag{14}$$

where $\mathbf{e}_n = [\ 1,\ \ e^{\frac{-j2\pi n}{N}},\ \ \cdots\ \ e^{\frac{-j2\pi(N-1)}{N}}\ ]^T$, $\mathbf{h}_{i,r} = [h_{i,r}[0], h_{i,r}[1], ...h_{i,r}[L-1], 0...0]^T$. The instantaneous SNR in terms of channel response can be expressed as [24]:

$$\lambda_{i,r}(n) = \left(\frac{E_s}{\sigma^2}\right) |H_{i,r}(n)|^2 \tag{15}$$

where E_s is the signal energy and σ^2 is the variance of zero mean complex white Gaussian noise. Applying (14) on (15), the following equation can be obtained:

$$\lambda_{i,r}(n) = \frac{1}{N}\left(\frac{E_s}{\sigma^2}\right) |\mathbf{e}_n^H \mathbf{h}_{i,r}|^2. \tag{16}$$

Using (12),(14) and (16), the total SNR at the output of MIMO-OFDM system for the n-th frequency bin is given by;

$$\lambda_{\mathrm{n}} = \frac{1}{N_T}\sum_{i=1}^{N_T}\sum_{r=1}^{N_R}\lambda_{i,r}(n), \tag{17}$$

Due to receive diversity within a MIMO receiver with the EGC, the SNR is given by [25];

$$\lambda_{\mathrm{equal}} = \lambda_{\mathrm{nd}}\left[1 + (N_R - 1)\frac{\pi}{4}\right], \tag{18}$$

where λ_{nd} is the output SNR without diversity i.e., for single input and N_R is the number of receive antennas. As EGC is applied to combine signals in different channels arising from the FS operation, the total SNR of MIMO-OFDM system with FS can be obtained in similar fashion in reference with baud rate sampling. The FS rate K can be considered as equivalent to the degree of receive diversity as in the above equation and received SNR in this case is the SNR of MIMO-OFDM system given by (17). Hence, the overall SNR for MIMO-OFDM system with FS can be written as:

$$\lambda_{\mathrm{FS}} = \lambda_{\mathrm{n}}\left[1 + (K - 1)\frac{\pi}{4}\right], \tag{19}$$

where λ_{n} is the SNR of MIMO-OFDM system with baud rate sampling given by (17).

IV. Proposed Iterative MIMO-OFDM System

There are three key elements, which have been used to explain the iterative detection process. They are: a priori (Δ), a posteriori (η) and extrinsic information (Γ) [26]. The priori information is defined as the information which is known before detection process starts and it generally arises from a source e.g., other component detector rather than the received sequence. The extrinsic information about a symbol u_k is the information provided by the detector based on both the received sequence and a priori information. While calculating extrinsic information, the detector excludes both the received samples representing systematic bits and priori information provided by another detector. The posteriori information about a symbol is defined as the information that the detector provides taking into account of all available sources of information about u_k. It is basically a posteriori Log Likelihood Ratio (LLR) that MAP algorithm provides as its output. The LLR output from a detector using MAP algorithm can be divided into three components as shown in following equation [26]:

$$\eta(u_k/\mathbf{y}) = \Delta(u_k) + L_c y_k + \Gamma(u_k), \tag{20}$$

where $\Delta(u_k)$ is a priori information provided by other decoder, $\mathbf{y}$ is the received symbol sequence [3], L_c is the channel reliability factor, y_k is the k-th received symbol and $\Gamma(u_k)$ is the extrinsic information provided by one detector to another detector.

Within the proposed iterative detection scheme as shown in Fig. 3, two different detectors detect the transmitted sequence from different fractionally sampled version of received signals. The MAP Algorithm is used in each detector stage and each detector gives the soft output in the form of the LLR value. The LLR value at particular sample instant is the ratio of the probability of the transmitted symbol being at one particular constellation. The MAP algorithm for each detected symbol gives the probability that the symbol to be within its predefined constellation, provided the received symbol sequence $\mathbf{y}$. This is equivalent to finding a posteriori log-likelihood ratio (LLR) $\eta(u_k/\mathbf{y})$ given by [27]:

$$\eta(u_k/\mathbf{y}) = \ln\left\{\frac{P(u_k = 1 \mid \mathbf{y})}{P(u_k = 0 \mid \mathbf{y})}\right\}, \tag{21}$$

In this work, the branch metric for MAP algorithm for an ISI channel has been modified so that the algorithm reduces the effect of channel by tracking the received sequence from forward as well as from backward. Branch metric reflects the transmitted symbol's constellation corresponding to particular received noisy sample at that sample instant. The value of branch metric is calculated differently depending on the calculation of probability for different constellations points. For an AWGN channel, the branch metric γ can be calculated by the following equation

[3]This is a symbol received sequence in the time domain as reflected in eqn. (5).

[26]:

$$\gamma_{t,t+1} = \prod_{i=1}^{n_0} \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left\{-\frac{1}{2}\left(\frac{y_{t,i} - u_{t,i}}{\sigma}\right)^2\right\}, \tag{22}$$

where $y_{t,i}$ is the i-th bit of t-th received symbol and $u_{t,i}$ is the i-th bit of branch symbol connecting states s_t and s_{t+1} in the trellis section, n_0 is the number of bits in a symbol. For simple Binary Phase Shift Keying (BPSK) transmission in ISI channel, the modified branch metric for input transmitted bit being $+1$ or -1 can be calculated as:

$$\gamma(\pm) = \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left\{-\frac{1}{2}\left(\frac{y_t - [\ \mp 1 \quad \mathbf{s}\ ]\mathbf{h}^H}{\sigma}\right)^2\right\}, \tag{23}$$

where $\mathbf{s}$ is a state vector whose dimension is one less than the dimension of the channel vector $\mathbf{h}$ as reflected in (23). For a 3-tap multi-path channel, $\mathbf{s}$ can have the following four possible values on the basis of channel memory: $[\ -1 \quad -1\ ]$, $[\ -1 \quad 1\ ]$, $[\ 1 \quad -1\ ]$ and $[\ 1 \quad 1\ ]$, and $\mathbf{h}$ is ISI channel vector e.g. $\mathbf{h} = [\ h_1 \quad h_2 \quad h_3\ ]$ for a 3-tap multi-path channel, represents the channel coefficients of different multi-paths.

The values of the LLR obtained by the above equation gives the complete soft information about the transmitted bit sequence. The sign of the LLR at each sample instant denotes whether the input information bit is $+1$ or -1 and magnitude of LLR denotes the probability of that particular symbol being in one particular constellation. The LLR values obtained in the above step are passed as soft input to the viterbi decoder in case of the coded system and in case of the uncoded system, a hard decision is made based on the sign of the LLR. In existing iterative system model in the literature, the iterative process is applied between a decoder and a detector. In this work, two fractionally sampled sequences are generated from the same sequence using fractional sampler and they are fed to two different detector stages. The decision is made based on the output of the second detector stage by passing it as the soft input to a hard limiting detector.

Let Δ_{ij} be the priori information for the i-th detector in the j-th iteration, Γ_{ij} is the extrinsic information provided by the i-th detector in the j-th iteration, η_{ij} is the complete or posteriori information provided by the i-th detector in the j-th iteration. In the first iteration, the extrinsic information to the first detector is taken as zero, therefore, the priori information to the first detector in the first iteration i.e.,Δ_{11} is also zero. The complete information η_{11} provided by the first detector in the first iteration can be written as:

$$\eta_{11} = \ln \left\{ \frac{P(u_k = 1 \mid \mathbf{y})}{P(u_k = -1 \mid \mathbf{y})} \right\}, \tag{24}$$

The extrinsic information from the first detector is calculated as:

$$\Gamma_{11} = \eta_{11} - \varsigma_1 - \Delta_{11}, \tag{25}$$

where ς_1 is the fractionally sampled version of the received signal with added AWGN noise fed to the input of the first detector, Γ_{11} is the extrinsic information provided by the first detector and Δ_{11} is the priori information to the detector which is taken as zero for the first iteration. Subsequently, Γ_{11} becomes priori information (Δ_{21}) for the second detector and the extrinsic information provided by second detector in the first iteration i.e., Γ_{21} can be written as:

$$\Gamma_{21} = \eta_{21} - \varsigma_2 - \Delta_{21}, \tag{26}$$

where ς_2 is a fractionally sampled version of the received sequence, fed to the second detector and η_{21} is the LLR values provided by the second detector in the first iteration. Γ_{21} is then used by the first detector in the second iteration. In this way, the process continues until the provided number of iterations is reached. In each iteration, the posteriori information value becomes more and more reliable than in the previous iteration.

V. Probability of BER for proposed Iterative MIMO-OFDM detector

In this section, we derive the probability of BER for the proposed iterative MIMO-OFDM receiver by referring to BER expressions for AWGN and Rayleigh fading channel carried out in [28]. For the purpose of completeness of this paper, we briefly include the steps for deriving these expressions from [28]. As shown in Fig. 3, the soft output from each detector can be written as: $\eta = \varsigma + \Delta + \Gamma$, where ς is the received discrete time sample to the particular detector, Δ is the priori information to the corresponding detector stage and Γ is the extrinsic information provided by that detector to the another stage. The detector output η is supposed to be distributed with mean value μ_η and variance σ_η^2 in case of AWGN channel. The bit error probability for BPSK modulated signal in AWGN channel can be written as [29]:

$$P_b = \frac{1}{2}\mathrm{erfc}\left(\sqrt{\frac{E_b}{N_0}}\right). \tag{27}$$

Assuming ς, Δ and Γ are independent to each-other,

$$\sigma_\eta^2 = \sigma_\varsigma^2 + \sigma_\Delta^2 + \sigma_\Gamma^2. \tag{28}$$

The soft output values from each stage detector can be written as:

$$\eta = ln\left(\frac{p(\mathbf{y}/u=+1)}{p(\mathbf{y}/u=-1)}\right). \tag{29}$$

The variance σ_η^2 can be written as [28]:

$$\sigma_\eta^2 = 8R\frac{E_b}{N_0}, \tag{30}$$

where R is the coding rate. We consider lower FS rate for our analysis and for lower FS rates, the correlation effect is negligible as reflected in our simulation result in Fig. 5. In this case, the fractionally sampled sequences can be considered to be uncorrelated. Then the priori information values Δ remain fairly uncorrelated from respective channel observations over many iterations. Also it can be assumed that the Probability Density Function (PDF) of extrinsic output values Γ approaches Gaussian-like distribution with increasing number of iterations [28]. Based on above two scenarios, the priori input Δ to the constituent detector can be modeled by applying an independent Gaussian random variable with variance σ_Δ^2 and mean zero in conjunction with the known transmitted systematic bit s.

$$\Delta = \mu_\Delta.s + n_\Delta, \tag{31}$$

where the mean value $\mu_\Delta = \sigma_\Delta^2/2$, and let S be a set containing all possible systematic information symbols i.e., $S \in (+1, -1)$ for the BPSK system. The conditional probability density function of Δ can then be written as:

$$p_\Delta(\xi/S=s) = \frac{e^{-\left((\xi-(\frac{\sigma_\Delta^2}{2}).s)^2\right)/2\sigma_\Delta^2}}{\sqrt{2\pi}\sigma_\Delta}. \tag{32}$$

The mutual information between the transmitted systematic bits S and the priori soft likelihood values Δ, $I_\Delta = I(S;\Delta)$ can be used to measure the information contents of a priori knowledge and can be written as in [28]:

$$I_\Delta = \frac{1}{2}\sum_{s=0,1}\int_{-\infty}^{\infty} p_\Delta(\xi/S=s) \times ld\frac{2.p_\Delta(\xi \mid S=s)}{p_\Delta(\xi \mid S=1) + p_\Delta(\xi \mid S=-1)}d\xi, \tag{33}$$

where $0 \leq I_\Delta \leq 1$. The above equation can be written as:

$$I_\Delta(\sigma_\Delta) = 1 - \int_{-\infty}^{\infty} \frac{e^{-((\xi-\sigma_\Delta^2/2)^2/(2\sigma_\Delta^2))}}{\sqrt{2\pi}\sigma_\Delta}.ld[1+e^{-\xi}]d\xi. \tag{34}$$

Above equation can be abbreviated as:

$$J(\sigma) = I_\Delta(\sigma_\Delta = \sigma) \tag{35}$$

with

$$\lim_{\sigma\to 0} J(\sigma) = 0, \ \lim_{\sigma\to\infty} J(\sigma) = 1, \quad \sigma > 0. \tag{36}$$

The variances σ_Δ^2 and σ_Γ^2 then can be written as [28]:

$$\sigma_\Delta^2 \approx J^{-1}(I_\Delta)^2, \sigma_\Gamma^2 \approx J^{-1}(I_\Gamma)^2, \tag{37}$$

where I_Δ is the mutual information for priori information and I_Γ is the mutual information for extrinsic information which can be obtained from Extrinsic Information Transfer Chart [28, 30, 31]. Applying the above equations, the expression for probability of BER for general iterative receiver in AWGN channel can be written as [28]:

$$p_b = \frac{1}{2}\text{erfc}\left(\frac{\sqrt{8R\frac{E_b}{N_0} + J^{-1}(I_\Delta)^2 + J^{-1}(I_\Gamma)^2}}{2\sqrt{2}}\right). \tag{38}$$

In Rayleigh fading channel, the bit error probability for BPSK modulated signal is given by [29];

$$p_b = \frac{1}{2}\left(1 - \sqrt{\frac{E_b/N_o}{(E_b/N_o)+1}}\right). \tag{39}$$

The conditional PDF of received signal y can be written as [28]:

$$p(y/S=s) = \frac{1}{\sqrt{2\pi\sigma_n^2}}\exp\left(-\frac{1}{2}\frac{(y-h.s)^2}{\sigma_n^2}\right). \tag{40}$$

From (38), let us define the factor G as:

$$G = 8R\frac{E_b}{N_o} + J^{-1}(I_\Delta)^2 + J^{-1}(I_\Gamma)^2. \tag{41}$$

For the case of proposed MIMO-OFDM with FS, the above equation can be modified by using (19) in the following form:

$$G_1 = 8R\lambda_n\left[1 + (K-1)\frac{\pi}{4}\right] + J^{-1}(I_\Delta)^2 + J^{-1}(I_\Gamma)^2, \tag{42}$$

where λ_n is the SNR of MIMO-OFDM system with baud rate sampling and K is the FS rate. Thus the probability of BER in an AWGN channel from eqn. (38) using factor G_1 can also be written as:

$$p_b = \frac{1}{2}\mathrm{erfc}\left(\frac{\sqrt{G_1}}{2\sqrt{2}}\right). \tag{43}$$

From equations (39), (42) and (43), the probability of BER for the iterative detection in a Rayleigh fading channel can be written in the following form since eqn. (39) is the derived expression in a Rayleigh channel provided the corresponding AWGN equation is (27):

$$p_b = \frac{1}{2}\left(1 - \sqrt{\frac{\frac{\sqrt{G_1}}{2\sqrt{2}}}{\frac{\sqrt{G_1}}{2\sqrt{2}} + 1}}\right). \tag{44}$$

VI. Simulation and Results

To evaluate the performance of the proposed systems, firstly, an OFDM system using FS has been simulated in AWGN and Rayleigh fading channels. The number of FFT sub-carriers considered is 64 and the total number of used data carriers out of these sub-carriers is 52. Cyclic prefix of 16 bit duration is considered and sampling frequency considered is 20 MHz. Hence, the cyclic prefix duration is of 0.8 μs and total data bit duration is of 3.2 μs. To reflect microwave transmission in wireless fading channels, modified Jakes channel model has been used to generate channel coefficients as in [32]. For simplicity, we use BPSK modulation scheme [4] at the transmitter. We have used MATLAB software as a tool for carrying out simulations.

Figure 4 show BER versus E_b/N_0 curves for the OFDM system without FS and the OFDM system with FS rates 2 and 3 times the baud rate for 2-tap and 5-tap Rayleigh fading channels [5]. For the 2-tap channel, there is approximately 2 dB E_b/N_o gain at BER value of 10^{-3} while going from the baud rate to the FS rate $K = 2$ and 1.5 dB of the same while going from $K = 2$ to $K = 3$. Similarly, for the 3-tap channel, almost 2 dB E_b/N_o gain is noted while increasing sampling rate from baud rate to $K = 2$ and 1 dB E_b/N_o gain while going from $K = 2$ to $K = 3$ at BER value of 10^{-3}. From the figure, it can be noted that there is gain in E_b/N_o while increasing the FS rate and the rate of increase varies for the channels having different taps.

[4]Higher modulation schemes can also be considered under the proposed framework by taking account of different signal constellations.

[5]We consider lower number of taps to reflect indoor picocell environment and higher number of taps can also be considered to reflect outdoor macrocell environment.

To observe the impact of spatial multiplexing along with the proposed receiver, a MIMO-OFDM system has been considered. Figure 5 shows the performance of the considered MIMO-OFDM system with baud rate sampling and with different FS rates. The FS rate is increased from 2 to 6 to observe the effect of FS rate on E_b/N_o gain improvement for a given BER. In this case, there is E_b/N_o gain of around 2 dB at BER of 10^{-3} in increasing the sampling rate from baud rate to $K = 2$ and gain of around 1 dB in increasing FS rate from $K = 2$ to $K = 3$. From the simulation results, it has been observed the increase in FS rate is in diminishing order as FS rate is increased and E_b/N_o gain after FS rate 6 becomes almost negligible. The diminishing order of E_b/N_o gain with FS rate is due to increased correlation of noise.

The information throughput enhancement obtained by using FS in different systems has been summarized in the Table I. This table shows the saving in E_b/N_o value (in dB) and incremental throughput (in bps/Hz) in simple OFDM system, (2×1) Alamouti system, (2×2) MIMO system and (2×2) MIMO-OFDM system for AWGN and flat fading Rayleigh fading channel. This table further shows information throughput values for FS rates of 2 and 3. The information throughput gradient and E_b/N_o gain are higher in case of FS rate 3 than in case of FS rate 2 for all the considered systems. However, increase of FS rate increases the complexity of receiver. Therefore, it can be claimed that the FS technique can enhance the throughput of MIMO-OFDM based systems at the expense of receiver complexity.

The comparative analysis for different combinations of velocities and data rates are provided in the tabular form in TABLE II. The carrier frequency of 2 GHz has been considered and velocities are taken as 5 km/h to 100 km/h. We consider that the speed 5 Km/h and 100 km/h to take into account of pedestrian and vehicular users. The data rates of 12.2 kb/s and 64 kb/s are considered to take into account of multiuser voice and data transmission. The effect of FS applied at the receiver may be different depending on the transmission data rates and velocities of the users. Therefore, we consider different combinations of transmission data rates and velocities to take into account of multiuser voice and data transmission for pedestrian and vehicular users. It can be observed from the table that doppler frequency is same for both voice (12.2 kb/s) and data (64 kb/s) transmission for the same velocity and channel length. Furthermore, it can be noted that the higher value of E_b/N_o is required at 100 km/h velocity and 12.2 kb/s for both

2-tap and 3-tap cases than in other cases.

To compare the BER performance of the proposed iterative structure with the performance of a non-iterative structure, a comparative simulation has been performed and the simulation result is shown in Fig. 6. In this simulation setting, a convolution code with $1/2$ code rate was used for both the iterative and non-iterative cases. The simulation parameters used for generating this result are presented in Table III. It can be observed that about 4.5 dB gain in E_b/N_o can be achieved at BER value of 10^{-2} using the proposed FS based iterative receiver. This E_b/N_o gain can be transferred to the enhancement in information throughput as: $C/B = \log_2(1{+}4.5) = 2.459$ bits per seconds per Hertz (bps/Hz). From this result, it can be concluded that the proposed FS based iterative detection scheme can significantly enhance the information throughput of MIMO-OFDM based systems.

VII. Conclusion

In this paper, a novel FS based iterative receiver has been proposed to enhance the information throughput of MIMO and OFDM based wireless communication systems. Firstly, the theoretical analysis of the considered MIMO-OFDM system has been carried out and the expression for the received SNR has been derived. Secondly, a novel iterative detector based on FS concept has been proposed and the theoretical expression for probability of BER has been derived. Finally, the performance of the proposed iterative receiver has been evaluated and compared with the non-iterative receiver in terms of BER performance. The simulation results show that the proposed MIMO-OFDM receiver with FS achieves significantly lower BER than the MIMO-OFDM with baud rate sampling. The increase in performance gain due to FS has been observed to be a non-linear function of the FS rate because of the noise correlation effect and it has been noted that the FS technique does not provide significant performance improvement after the FS rate of 6. It can be concluded that the proposed FS based iterative receiver can provide significantly better performance than non-iterative detection techniques in wireless fading channels. We consider the convergence behavior analysis of the proposed receiver using the EXIT chart as well as the implementation of this concept for higher modulation schemes and for higher FS rates as our future work.

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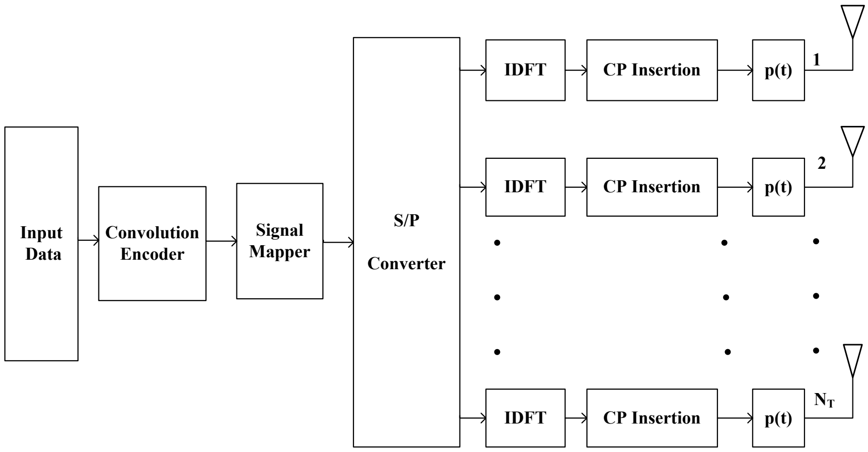

Fig. 1: System model for MIMO-OFDM Transmitter

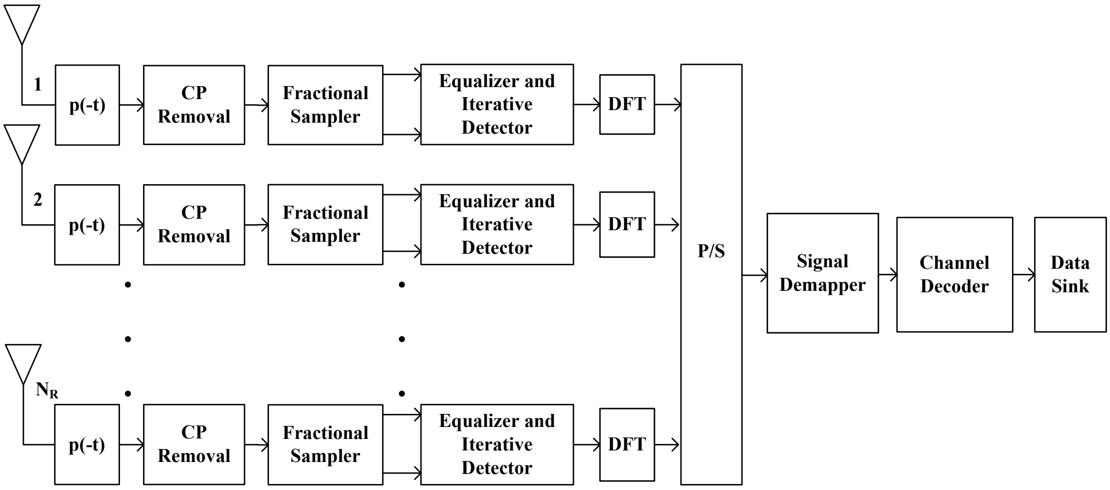

Fig. 2: System Model for proposed Iterative MIMO-OFDM Receiver

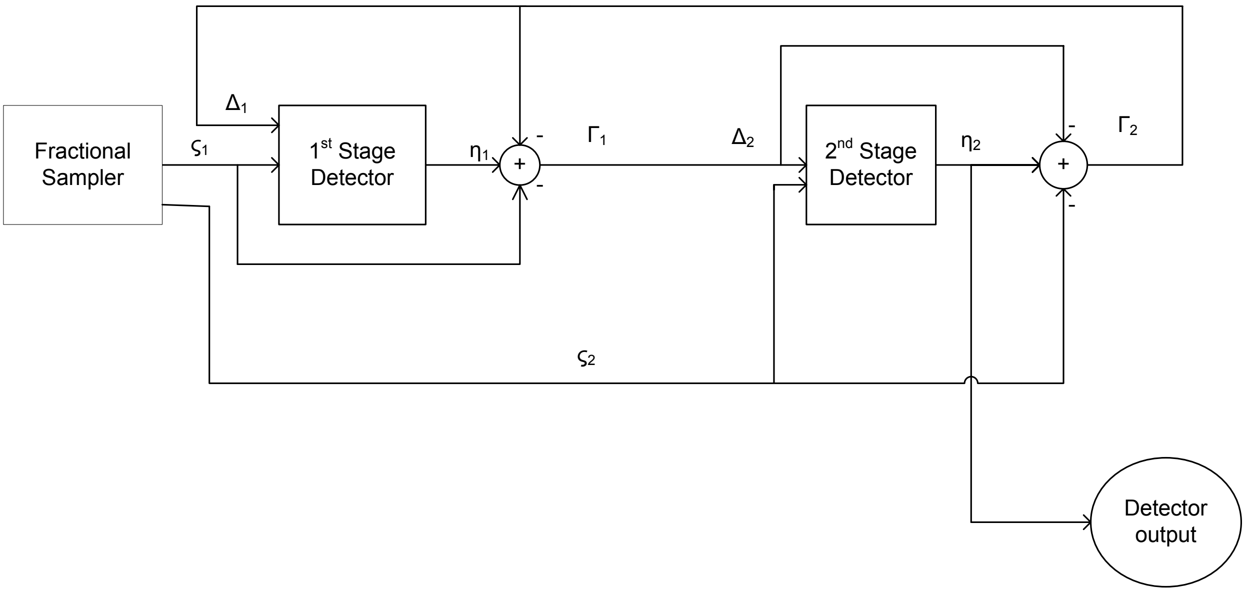

Fig. 3: Simplified Process diagram of the proposed iterative detection from FS signal sequence

TABLE I: Comparison of performance improvement due to FS for different systems

Systems	FS Rate=2				FS Rate=3			
	AWGN		Rayleigh		AWGN		Rayleigh	
	Saving in E_b/N_o (dB)	Incremental Throughput, (bps/Hz)	Saving in E_b/N_o (dB)	Incremental Throughput (bps/Hz)	Saving in E_b/N_o (dB)	Incremental Throughput (bps/Hz)	Saving in E_b/N_o (dB)	Incremental Throughput (bps/Hz)
OFDM	3 .0	2.0	2.5	1.8075	4.8	2.5361	4.5	2.4594
Alamouti (2tx, 1rx)	3 .0	2.0	2.0	1.5850	4.8	2.5361	3.5	2.1699
MIMO (2tx, 2rx)	3.0	2.0	1.5	1.3219	5.0	2.5850	2.5	1.8074
MIMO-OFDM	1.5 dB	1.3219	1.0	1.0	2.5	1.8074	1.5	1.3219

TABLE II: Comparison of BER performance of MIMO iterative detector for different scenarios

Channel length	2-tap				3-tap			
Velocity (Km/h)	5		100		5		100	
Data Rate (Kb/s)	12.2	64	12.2	64	12.2	64	12.2	64
E_b/N_o (dB) value at BER of 10^{-3}	4.85	4.3	5.2	4.5	3.8	3.6	5.2	4.85
Fading Rate	7.589e-04	1.446e-04	0.0152	0.0029	7.589e-04	1.446e-04	0.0152	0.0029
Doppler frequency (Hz)	9.2593	9.2593	185.18	185.18	9.2593	9.2593	185.18	185.18

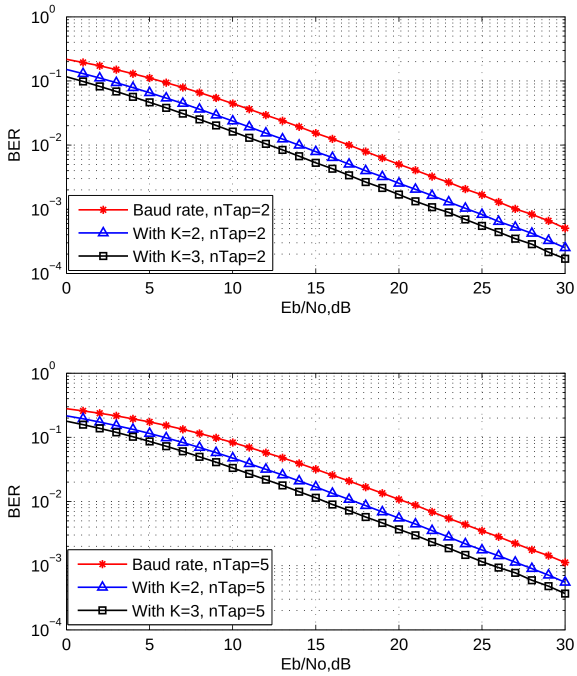

Fig. 4: BER versus Eb/No for OFDM in 2-tap and 5-tap Rayleigh channel

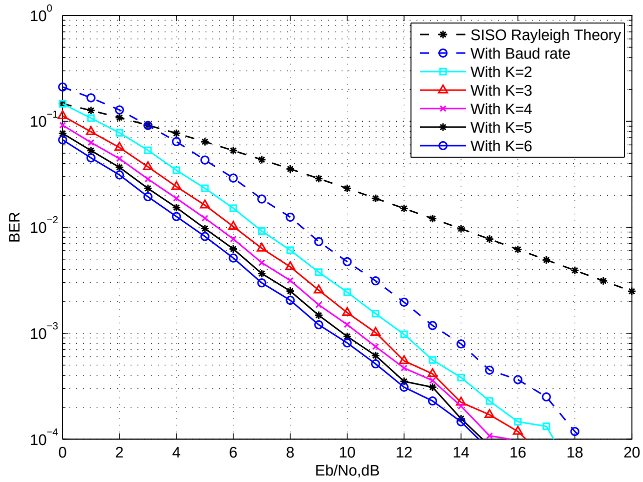

Fig. 5: BER versus Eb/No for FS-MIMO-OFDM (2×2) in Rayleigh channel

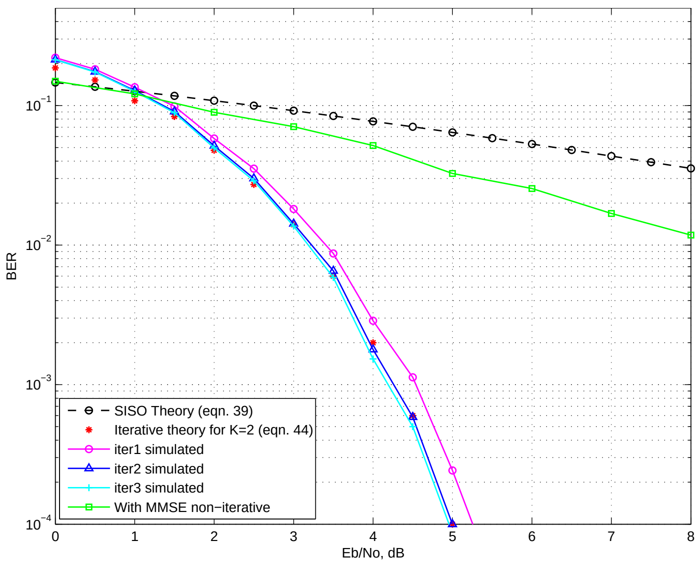

Fig. 6: BER performance comparison of proposed iterative receiver and non-iterative MMSE method

TABLE III: Simulation Parameters for Fig. 6

Parameters	**Values**
Channel	Multipath Rayleigh fading channel
Number of Taps	3
Modulation	BPSK
Channel Encoder	Convolution encoder
Coding Rate	1/2
Decoder	Viterbi
Memory length	2
Number of Trellis states	4
Puncturing	No
Number of decoders	2 (Fig. 3)