Adaptive Code Synchronisation with Reduced Computational Complexity and Premodulation Filter for Improved Capacity in MIMO System

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Abstract
A wireless channel model can be considered both as a time-domain and frequency-domain channel with stationary and ergodic features. As a time-domain channel it is complex normal and independent across filter taps and users. Similarly in frequency-domain it has correlated channel gains across subcarriers. In this context, in this paper, anticipative models where decisions are made based on the distribution and the actual realizations of the random quantities (adaptive models) are presented for MIMO system. A novel premodulation filter and adaptive code synchronization scheme is discussed. Multistage recourse models with non-anticipative action for the 1st stage and recourse actions for the 2nd stage based on the realization of the random quantities are presented. The work shall assist in improved definitive prioritization among users and can trace boundary of capacity region with specified ratio. The proposed algorithms are adaptive with low complexity and more suited for online processing.

Keywords
MIMO, Channel fading, Pre-modulation filter, Adaptive Code Synchronization.

I. Introduction

The algorithms and physical modules presented in this work shall assist in forming a unified framework for optimal resource allocation in a MIMO system. The work performs a granular exploitation of diversity among users through channel state information (CSI) feedback. Previous related works have achieved some form of temporal diversity by maximizing an exponentially windowed running average of the rate. However, the reported works have not exploited the time varying nature of wireless channel. Also, most of the previous works assume perfect channel state estimation

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And these are unrealistic due to channel estimation errors and delay. However, the present work can tolerate imperfect channel knowledge and the allocate is based on statistics of Channel estimation/prediction errors. This paper reports capacity based analysis and performs comparison study previous works assume perfect channel estimation and these are unrealistic due to channel estimation errors and delay. However, the present work can tolerate imperfect channel knowledge and the allocate is based on statistics of Channel estimation/prediction errors. This paper reports capacity based analysis and performs comparison study with conventional MIMO transmission schemes. The rest of the paper is organized as follows: Section II to section IV describes briefly about channel fading effects, adjacent channel interference and factors that contribute to spectrum/frequency reuse. Section V describes Digital modulation schemes and design of premodulation filter. Section VI describes the channel estimation model. Section VII explains the transceiver design and a novel adaptive code synchronization circuit is described in section VIII of this paper. The design of MIMO system

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using a complex version of the original Alamouti Code is described in Section IX of the paper. Section X explains the results and discussion and shows the improvements in capacity and reduced computational complexity achieved with the present scheme. The conclusions and future direction of research are given in section XI.

II. Adjacent Channel Interference (ACI)

The ACI is defined as ACI = pSL/PML,,, where pSL is the power spilling over in to the channel of interest due to the side lobes of an adjacent sub channel and PML is the power produced in that channel due to its own main lobe . Assuming the main lobe occupies the band $-\frac{1}{T} \le f \le \frac{1}{T}$, the adjacent band on the left, lies inside the side lobe $-\frac{3}{T} \le f \le -\frac{1}{T}$, similarly the adjacent band on the right lies inside the side lobe $\frac{1}{T} \le f \le \frac{3}{T}$. Hence, ACI can be defined as

ACI=

$$\frac{\int_{-\frac{1}{T}}^{\frac{1}{T}} |p\, f|^2\, df}{\int_{-\frac{1}{T}}^{\frac{1}{T}} |p\, f|^2\, df} = \frac{\int_{1/T}^{3/T} sin2\,(fT)df}{\int_{-1/T}^{1/T} sin2\,(fT)df}$$

In general, ACI is defined as,

$$\text{ACI}(\delta f) = \frac{\int_{-\infty}^{\infty} G(f)\,|H\,(f-\partial f)|^2\, df}{\int_{\infty}^{\infty} G\,(f)\,|H\,(f-\partial f)|^2\, df}$$

Where, G(f) -PSD of i/p signal H(f)- freq response of BPF used to separate adjacent channel Δf - frequency separation between channels. It is evident, that the filter should be selective enough for the main lobe of G(f) to lie inside the pass band of H(f).

III. Reducing fading effects

An interleaver is typically used over fading channels to boost the performance of forward error correction codes. An interleaver takes a block of data from the encoder and permutes the order of the bits before transmitting them. The ideal spacing of bits in an interleaver, so as to achieve the objective of making the fading on adjacent bits appear uncorrelated, is given by

$$\tau = \frac{-2.2C}{2\pi f_e V}$$

Where, τ =Decorrelation time

$C=3\times10^{8}$ m/s

f_e=Transmitting frequency

V= Velocity of a mobile terminal

A. Pilot symbol to track fading and residual frequency errors

In this research , the use of pilot symbols is made to track both fading and small residual errors in the receiver .This is possible , since for known pilot symbols, the noisy estimate of the fading frequency error can be expressed as

$$h_{ki} = \alpha_{ki}\, e^{(2\pi\Delta kT+\theta)} + w_{ki}$$

(1)

B. Interleaver to reduce fading

Two types of interferences are reported namely, block and convolution interleavers. The comparison between them is given in table 1

Table 1 Comparison among interleaver schemes

Parameter	Block interleaver of size LXN	Convolution Interleaver of size LXN
End-to –end delay	2LN	1(N-1)
Memory requirement	LN	½ L (N-1)
Commutator synchronization	Simple	Complex
Computational complexity	Complex	Simple

The Rayleigh fading model used for study in this work is shown in Figure 1.

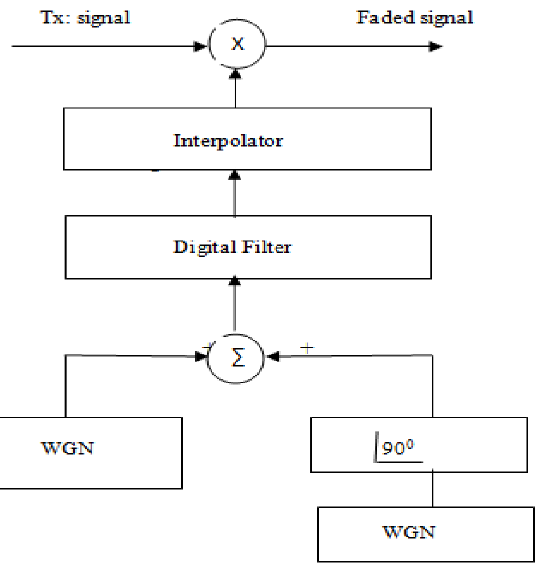


Fig. 1 Digital model for simulating Fast Rayleigh fading channel

In Figure 1 to achieve fast fading, the spectrum shaping filter needs to be very narrow band and this requires many interpolation stages.

IV. Factors Influencing Spectrum usage/Reuse

Table 2 lists the factors that are mainly responsible for the increase in the spectrum usage. Many of the factors listed in the Table-2 permit spectral reuse (Frequency reuse), Temporal reuse (Spectrum sharing) and Spatial reuse (Spectrum reuse). When Fast Rayleigh fading exists, it is too quick to be compensated by power control, Hence, inter leaving and forward error correction coding are the alternative solutions used. Also, in intercell, interference, the residual variations are effectively handled by interleaving.

Table 2 Factors influencing spectrum usage/reuse

Factors	Remarks
Antenna Gain	Improved directionality means improved frequency reuse.
Modulation	Increased Spectral efficiency makes the spectrum more compact and reduces ACI. Effective modulation schemes concentrate on generating optimal power to reduce interference level or alternately for a given transmit power, the minimum separation among mobile units to avoid interference can be determined.
Filters	Match filtering improvements permits steeper band roll offs and a more compact spectral shape.
Oscillators	Spectrum becomes compact with improved stability and accuracy of oscillators. If the accuracy of oscillator is less, then increase phase rotation will occur during one symbol period. Also, there shall be residual frequency error dependent on the relative error. Thus, the maximum frequency error imposes a constraint on the maximum permissible phase rotation between two successive symbols.
Semiconductor Technology	Improved switching speeds and good immunity to RFI and EMI. Improved Circuit designs result in reduced susceptibility to interference.
Digital Processing	Improved algorithms to track phase and gain with great efficiency resulting in low implementation losses. Also, there will be an increase in the amount of frequency reuse.
Coding	Uses of error correcting codes allow large improvements in Spectral efficiency. Typically, receivers that use error coding can operated lower power and tolerate more interference.

V. Digital Modulator

The digital modulator consists of pre-modulation filter required to produce a data-dependent phase signal $\theta(t)$.The terms $\cos[\theta(t)]$ and $\sin[\theta(t)]$ can be obtained using two read-only-memory (ROM) units. The resultant digital signal is converted into analog form. This model can effectively deal with non linearities. The constraint is that the height frequency component of both $\cos[\theta(t)]$ and $\sin[\theta(t)] < fc$. This effectively defines the bandwidth of $\cos[\theta(t)]$ and $\sin[\theta(t)]$ and larger the bandwidth demand, the higher is the significant frequency component (Figure 2).

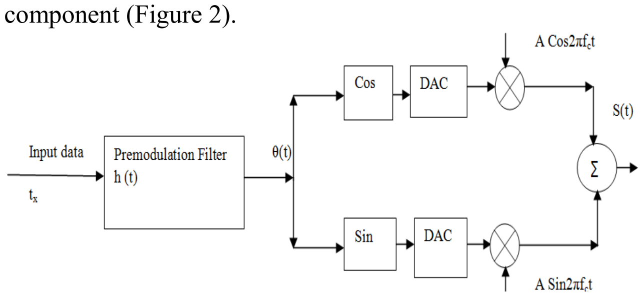


Fig. 2 Proposed Digital modulator with the premodulation filter

A. Pre-modulation Filter

In this work, the pre-modulation filter is implemented using tapped –delay-line filter. Thus,

$$\theta(t) = \sum_{k=0}^{N-1} h_k\, b_k(t - kT) \quad \ldots (2)$$

(or)

$$\theta(t) = \int_0^{\infty} h(\tau) b_k (t - \tau) d\tau \quad \ldots (3)$$

Where τ = KT and T is the Sampling interval.
For Small values of "T", eqn. (2) can be approximated as

$$\theta(t) = \sum_{k=0}^{\infty} h\ (kT)\, b_k (t - kT) . T \quad \ldots (4)$$

Eqn. (4) is effectively a tapped delay line filter relation and the pre-modulation filter is shown in fig.3

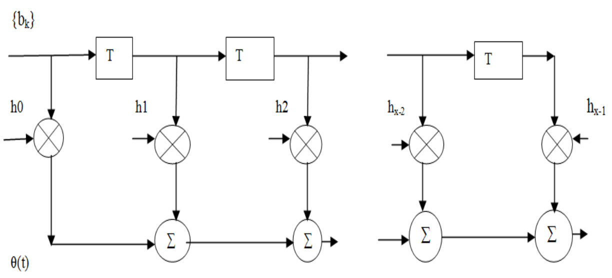


Fig. 3 Tapped delay line filter implementation of premodulation filter

B. Modulation for wireless channel

In this paper, the 16- PSK modulation scheme is preferred compared to a 16-QAM scheme. The primary reason is that, 16-QAM is hybrid modulation scheme and for it to work properly on a wireless channel the channel should be necessarily linear, else received signal will be distorted. But, the 16-PSK scheme will be impervious to nonlinear distortion among the PSK schemes, the major differences between Quadriphase shift keying and minimum-shift keying for digital transmission of binary data over a wireless channel is listed in table 3. To further enhance the advantage of MSK, a Gaussian prefilter can also be used [GMSK].

Table 3 Comparison among modulation schemes for different parameters

Parameter	QPSK	MSK
Computational Complexity	Simple, since linear modulation	Complex, since nonlinear modulation
Spectral occupancy	The base band PSD α $1/f^2$ Interference outside the signal band of interest is more.	The base band PSD α$1/f^4$ The interference outside the signal band of interest is less.

VI. Channel Estimation

In this paper, the impulse response of the wireless communication channel is estimated by using a sounding sequence. The sounding sequence chosen is a long PN sequence, with closer auto correlation function. The tradeoff is that the synchronization delay is slightly more and also the computational complexity is higher. This is shown in figure 4. Alternatively, the channel state information can also obtained by impulse training sequence in each transmitted packet. However this could occupy a portion of each packet.

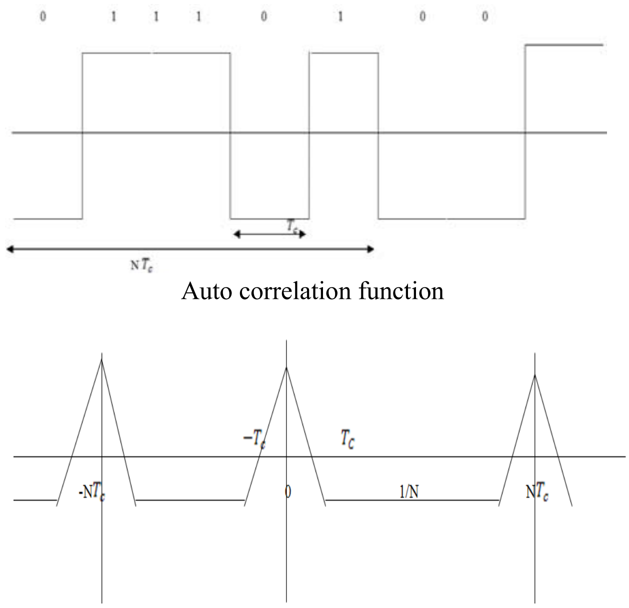


Fig. 4 PN sounding sequence and its ACF, Note: T_c --- duration of symbol, N------ Length of PN sequence

In this paper, the pilot channel is typically used for channel estimation. The correlator output at the synchronized time provides the channel estimate at the zero chip offset. Similarly the correlator output at the one chip offset provides the channel estimate for the second tap and so on for other taps. Thus, continuous tracking of the pilot channel can track the frequency error and other changes in the channel. The novelty of the approach is that, the pilot channel need not be modulated and hence, there is no effect of decision errors. Consequently, the technique can work at reduce SNRs. Also there is no delay tracking the channel. The channel estimation model proposed in this work helps to improve the capacity gain of MIMO wireless system with SNR improvement independent of channel errors.

VII. Transceiver Module

The channel encoder used is a two stage serial turbo encoder and is shown in figure 5. The multiplexer in the outer encoder is used to serially combine the message bits, parity-check bits I and Z into a single serialized binary stream. The receiver model proposed in this work is a novel iterative network and decodes the channel output in response to the encoder. The proposed receiver (figure 6) has the following features of (i) Computationally more complex, since it is a multi-stage feedback system (ii) However, performance is improved and approaches close to Shannon limit. Thus, the proposed receiver provides trade off of increased computation complexity for improved receiver performance.

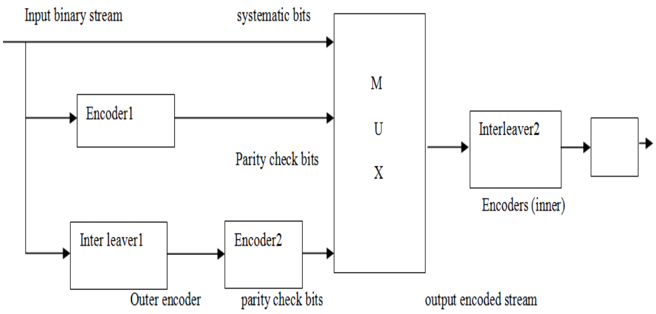
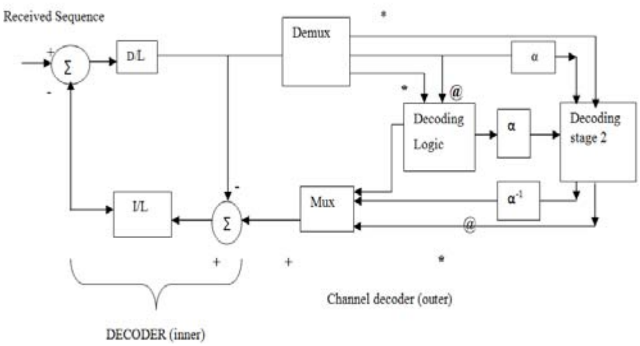


Fig. 5 A two stage serial turbo channel encoder



Note: D/L - De-interleaver, I/L - interleaver, * - Soft decoded parity bits, @ -Soft decoded info bits

Fig. 6 Proposed Iterative receiver

Conventional equalization decoding receivers (even though closed loop feedback system) suffers from the drawbacks, that the receivers may become unstable. This is because, the deinterleaver –channel decoder-interleaver chain feeds back into the equalizer a stream of parity-check bits that are more accurate than those contained in the received signal, which, intern, tends to make the equalizer more accurate. Thus, the closed loop system is liable to become unstable at lower SNR. At lower SNR, the incoming noisy parity- check bits become more unreliable causing the iterative receiver to diverge and thereby create increased number of errors.

VIII. Code Synchronization

The structure of the code synchronization circuit used in this work is novel and can handle a stream of complex base band samples. The circuit separate the real and imaginary processing and can adapt itself when

(i) Modulation is QPSK and the spreading code is same in both I and Q channels.
(ii) Modulation is QPSK and spreading code is different in each of the I and Q channels.

To ensure the adaptive behavior of the code synchronization circuit, an additional term (2nd term) is added in the correlator output to obtain eqn. (5)

$$R = \sum_{K=1}^{K1} C_i\,(KT) + jC_q\,(KT))\,(r_i\,(KT) + jr_q\,(KT))$$

$$K1 = \sum_{K=1} C_i(KT)(r_i(KT)+jr_q(KT)) + jC_q(KT)\,(r_i(KT)+jr_q(KT)) \quad \text{----------}\ (5)$$

The adaptive nature of code synchronization circuit allows the inclusion of cross terms between I and Q. The adaptive code synchronization circuit is shown in figure 7. The following reasonable assumptions are made in the implementation of the code synchronization circuit:

(i) Small frequency error and hence minimal phase rotation occur
(ii) Data modulation does not occur
(iii) Minimal noise

The above assumptions can also be achieved by using sectored antennas as these will effectively reduce the interference caused by signals generated at a base station, since signals intended for other users is no longer broadcast in all directions.

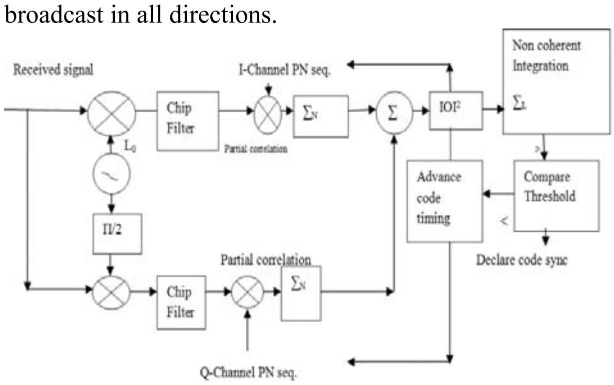


Fig. 7 Adaptive Code Synchronization with augmented module to include cross terms between I and Q

IX. The complete MIMO receiver

The MIMO receiver for two antennas on transmit and receive, including the respective additive noise components and the channel estimation model discussed in section 6 of this paper is shown in Appendix-I. The channels between the transmit and receive antennas and outputs of the receive antennas at times t and t +T is shown in Table 4a and Table 4b

Table 4a channel model between Tx: and Rx: antennas

	Rx: antenna1	Rx:antenna2
Tx:antenna1	h1	h3
TX: antenna2	h2	h4

Table 4b outputs of receive antennas

	Rx: antenna1	Rx:antenna2
Time $t^{\mid}$	$\tilde{x}_1$	$\tilde{x}_3$
Time $t^{\mid}$+T	$\tilde{x}_2$	$\tilde{x}_4$

A. Expression for receive antenna outputs

$\tilde{x}_1 = h_1\tilde{s}_1 + h_2\tilde{s}_2 + \tilde{w}_1$ at time $t^{\mid}$ … (6)

$\tilde{x}_2 = -h_1\tilde{s}_2^* + h_2\tilde{s}_1^* + \tilde{w}_2$ at time $t^{\mid}$+T … (7)

$\tilde{x}_3 = h_3\tilde{s}_1 + h_4\tilde{s}_2 + \tilde{w}_3$ at time $t^{\mid}$ … (8)

$\tilde{x}_4 = -h_3\tilde{s}_2^* + h_4\tilde{s}_1^* + \tilde{w}_4$ at time $t^{\mid}$+T … (9)

The linear combiner outputs in terms of the receive antennas output and the channel impulse response is given by

$$\tilde{y}_1 = h_1^*(h_1\tilde{s}_1 + h_2\tilde{s}_2 + \tilde{w}_1) + h_2(-h_1\tilde{s}_2^* + h_2\tilde{s}_1^* + \tilde{w}_2)^* + \tilde{h}_3^*(h_3\tilde{s}_1 + h_4\tilde{s}_2 + \tilde{w}_3) + h_4(-h_3\tilde{s}_2^* + h_4\tilde{s}_1^* + \tilde{w}_4)$$

$$=(\alpha_1^2 + \alpha_2^2 + \alpha_3^2 + \alpha_4^2)\tilde{s}_1 + \tilde{v}_1$$

where,

$h_k = \alpha_4 e^{j\varphi k}$ for all k=1,2,3,4 and

$\tilde{v}_1 = h_1^*\tilde{w}_1 - h_2\tilde{w}_2^* + h_3^*\tilde{w}_3 - h_4\tilde{w}_4^*$

Similarly,

$\tilde{y}_2 = (\alpha_1^2 + \alpha_2^2 + \alpha_3^2 + \alpha_4^2)\tilde{s}_1 + \tilde{v}_2$

where, $\tilde{v}_2 = h_2^*\tilde{w}_1 + h_1\tilde{w}_2^* + h_4^*\tilde{w}_3 + h_3\tilde{w}_4^*$

In the above $\tilde{v}_1, \tilde{v}_2, and\ \tilde{w}_k$ are complex Gaussian-distributed with zero mean and common variance σ_0^2

B. Decision rule for symbols $\tilde{s}_1$and $\tilde{s}_2$

$\hat{s}_1$ = estimate of $\tilde{s}_1$ = arg min $\{d^2(\tilde{y}_1, (\alpha_1^2 + \alpha_2^2 + \alpha_3^2 + \alpha_4^2)\phi\}$
$\phi \in S$

$\hat{s}_2$ = estimate of $\tilde{s}_2$ = arg min $\{d^2(\tilde{y}_2, (\alpha_1^2 + \alpha_2^2 + \alpha_3^2 + \alpha_4^2)\phi\}$
$\phi \in S$

X. Results and Discussion

Figures 8 to figure 11 shows the input data stream through a fast fading channel and recovered with the proposed receiver (figure 6). The performance of the designed system is evaluated for different types of model such as (i) white Gaussian Noise (ii) Rayleigh Fading and (iii) Rician Fading. It should be noted that frequency shifting and phase rotation do not affect the statistical properties of Zero-mean WGN. The capacity plot and the computational complexity for the proposed receiver as the number of users vary is obtained and shown in figure 12 and figure 13. The obtained results clearly demonstrate the effectiveness of the proposed scheme both in terms of improved capacity and reduced computational complexity. This is due to the reason that, in this work, the code 'S' representing the original Alamouti code is chosen as complex, as compared to a real valued representation. Thus, matrix multiplications are less intensive compared to the real version.

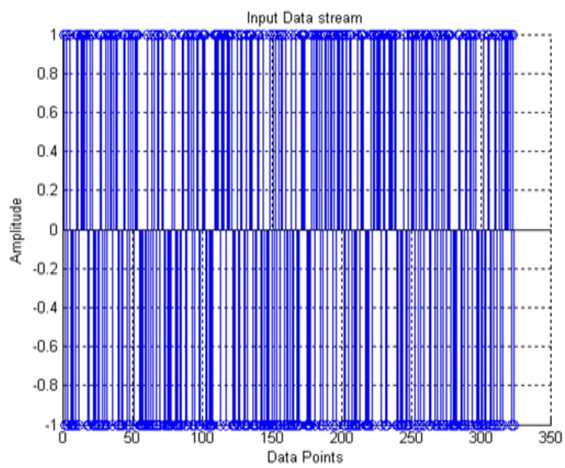


Fig. 8 Input data stream

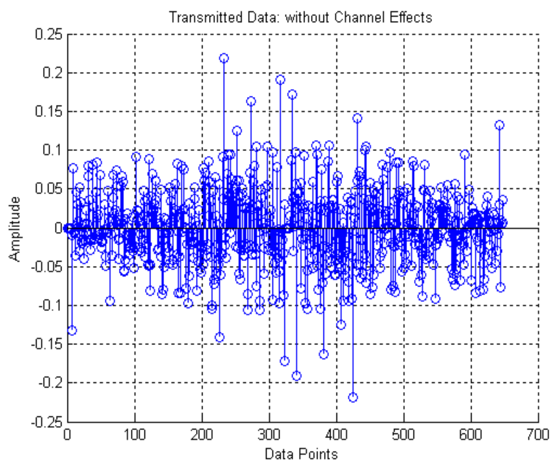


Fig. 9 Transmitted data without channel effects

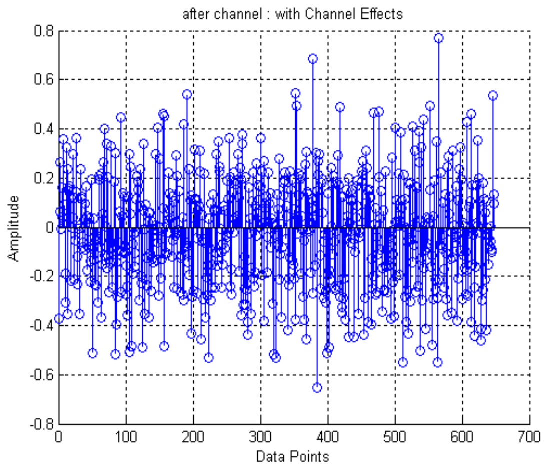


Fig. 10 Transmitted data with channel effects

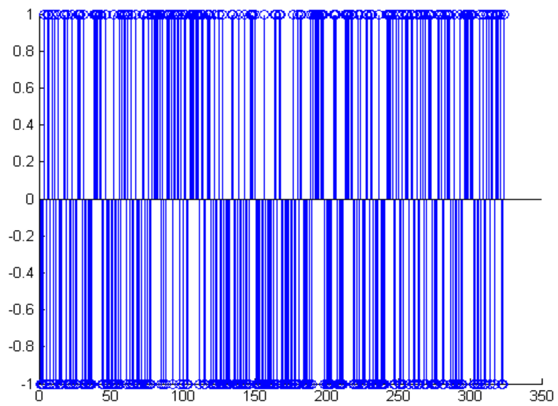


Fig. 11 Demodulated output i.e. decoded data

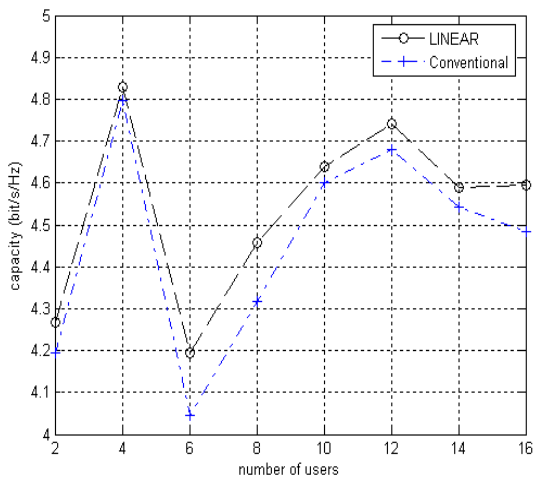


Fig. 12 Capacity Vs number of users graph for the linear (i.e. proposed) receiver

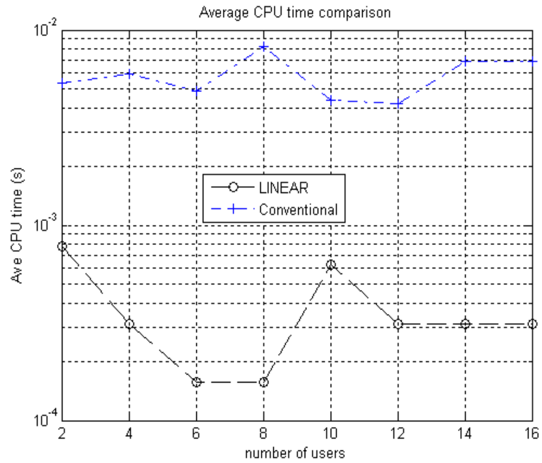


Fig. 13 Computational complexity graph for the proposed (i.e. linear) receiver

XI. Conclusion

In this paper, a maximum likelihood decoder based MIMO scheme was presented. To reduce the computational complexity a complex version of the Alamouti code was used. The channel capacity of the code is bounded by the sum of two single-input, single-output capacities. It is also shown that the receive diversity order is expandable by increasing the number of receive antennas. Also, the proposed MIMO receiver is attractive, in that, it requires a linear model only. The proposed work has several novel features such as (i) the use of a channel estimation model without the need to demodulate the channel and (ii) a computationally less complex code for the original Alamouti code. Future direction of study shall focus on the study of whether the obtained results are scalable for distributed mode of usage also.

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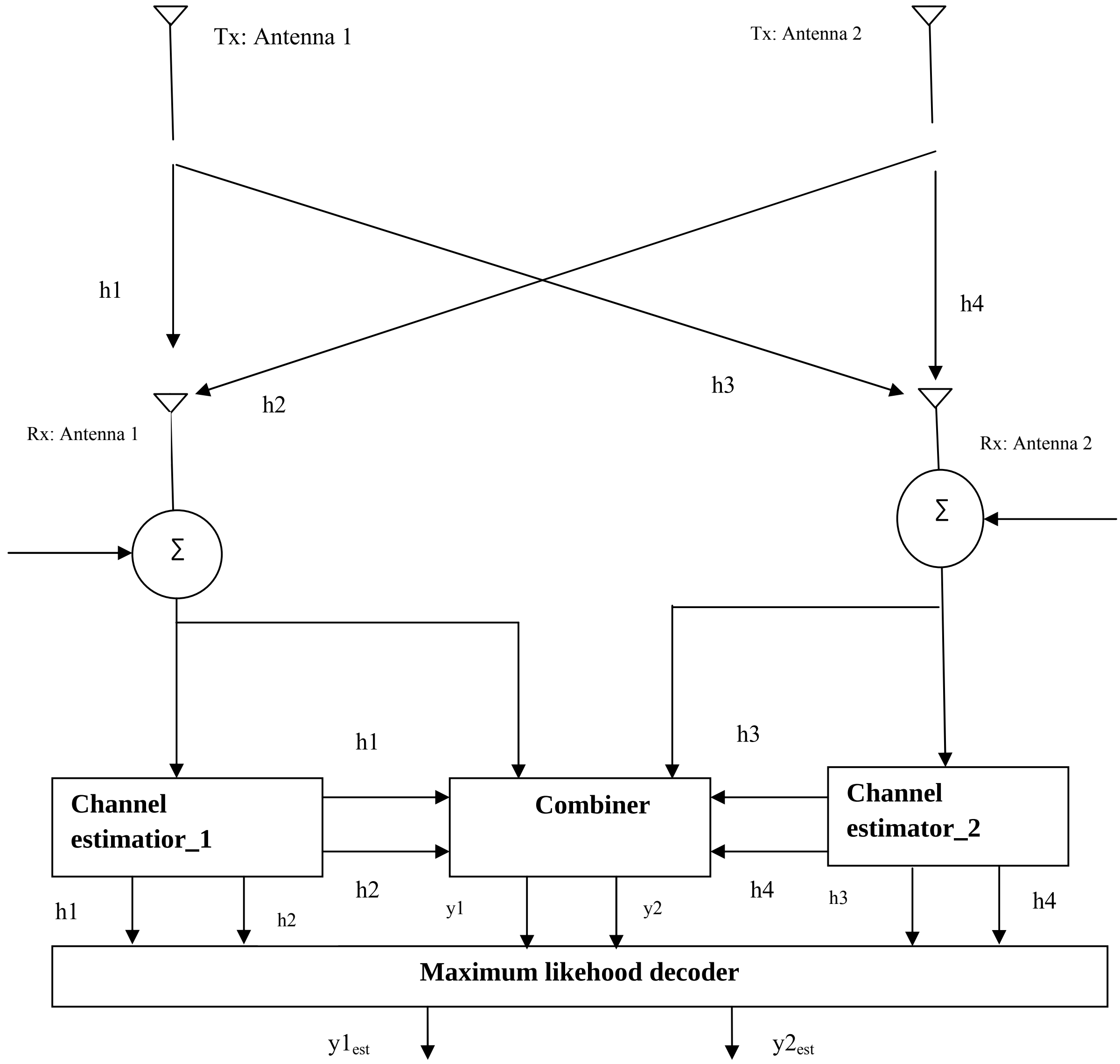
APPENDIX – I


Tx: Antenna 1
Tx: Antenna 2
h1
h4
h2
h3
Rx: Antenna 1
Rx: Antenna 2
Σ
Σ
h1
h3
Channel estimatior_1
Combiner
Channel estimator_2
h2
h4
h1
h2
y1
y2
h3
h4
Maximum likehood decoder
y1est
y2est