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Iterative-Equalization for Single-Carrier Transmission over Adaptive Multiuser Receivers Scheme for MIMO OFDM

Lokesh N S and Dr Suresh H N

Research Scholar, Jain University, Bangalore
Professor, Bangalore Institute of Technology, Bangalore
Email: lokgwd20@gmail.com, hn.suresh@rediffmail.com
Corresponding Author: Lokesh N S

Abstract

Adaptive multiuser receivers scheme for MIMO OFDM over Iterative-Equalization for Single-Carrier Transmission, which we refer to as Iterative AMUD MIMO OFDM. It involves the joint iteration of the adaptive minimum mean square error multiuser detection and decoding algorithm with prior information of the channel and interference cancelation in the spatial domain. A partially filtered gradient LMS (Adaptive) algorithm is also applied to improve the convergence speed and tracking ability of the adaptive detectors with slight increase in complexity. The proposed technique is analyzed in slow and fast Rayleigh fading channels in MIMO OFDM systems. The Adaptive Multiuser Detection for MIMO OFDM system (AMUD MIMO OFDM) performs as well as the iterative equalization for single-carrier for higher modulation scheme. The LMS algorithm and maximum a posterior (MAP) algorithm are utilized in the receiver structures. The results of the higher modulation schemes shows that as the modulation order increases, a higher SNR is required to obtain the same BER performance at a lower order. The iterative gain is greater for higher modulation orders as compared with lower modulation

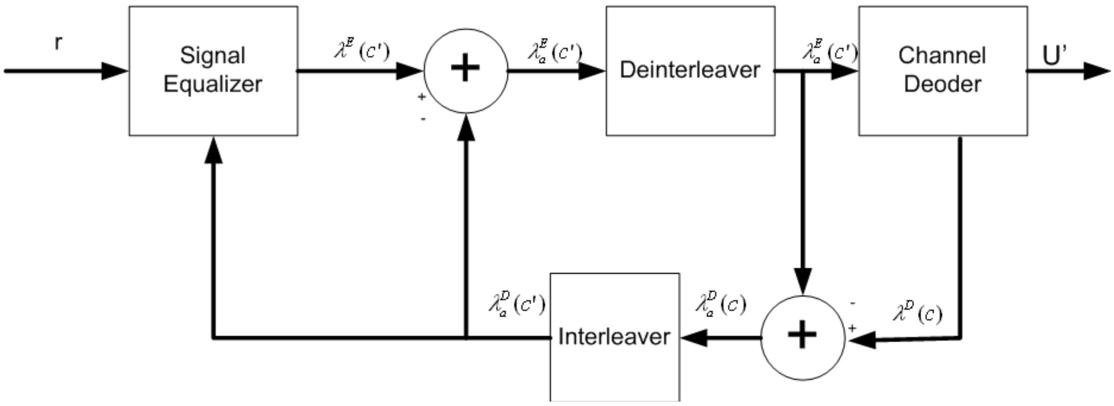
Keywords: MIMO OFDM, Adaptive filter, Turbo codes, Iterative decoding

INTRODUCTION

Wireless Local Area Network standards employ OFDM, which offers high spectral efficiency and superior tolerance to multi-path fading (IEEE 2003). In OFDM computationally-efficient Fast Fourier Transform (Cimini 1985) is used to transmit data in parallel over a large number of orthogonal subcarrier which is maintained even in frequency selective fading (Zelt 2000), (Batariere 2001), (Y. Li 1999). Throughput and capacity can be improved when multiple antennas are applied at the transmitter and receiver side, especially in a rich scattering environment (Foschini 1998), (Raleigh 1998) as well as frequency-selective fading channels.

The conventional approach implements an equalizer to remove ISI or use MAP or maximum likelihood (ML) detection. Data reliability can be enhanced using coding, where the data is encoded in the transmitter prior to transmission. For reasons of complexity, the receiver then typically performs separate equalization and decoding of the data. Significant performance gains can be achieved through joint equalization and decoding at the cost of added complexity. A recent approach that significantly reduces the complexity of joint equalization and decoding is called ”turbo equalization” algorithm, where MAP/ML detection and decoding are performed iteratively on the same set of received data. It has recently been stated that passing soft information, the use of interleaving, and the controlled feedback of soft information are essential requirements to achieve performance gains with an iterative system (Bauch 1998), (Tuchler 1998). The iterative principle has been extended to encompass single carrier equalization techniques, this allows single carrier systems to combine the operations of equalization and channel coding to operate in a wideband channel with performance that could not previously be achieved with traditional equalization and forward error correcting (FEC) techniques (Douillard 1995). Iterative equalization techniques have been shown to give excellent error rate performance for both fixed and fast fading channels (Bauch 1995).

Adaptive MMSE Multiuser detection (AMUD) is for demodulation of digitally modulated signals with multiple access interferences (MAI). Conventionally, individual channel estimation as stated by (Teletar 1999) was improved by joint estimation as stated in (Eneh 2010). This scheme was designed for total elimination of MAI in the system. In a single user environment, every match filter maximum likelihood receiver plays the role of Adaptive MMSE maximum likelihood receiver (Rapajic 1994), (Rapajic 1999). In the implementation Adaptive Minimum Mean Square Error Multiuser Detection (AMUD), provides robustness and mobility in a time variable frequency selective multipath fading channel, it improves the bit error rate performance and therefore enhances channel capacity of a multi-cellular environment. MIMO OFDM mitigates multiple access interference and increases capacity (Sampath 2002), (St¨uber 2004). In (Rapajic 1999) A MMSE MUD techniques was used effectively to achieve the performance of a maximum likelihood estimator but on a linear complexity.



The contributions in this paper, AMUD MIMO OFDM over Iterative-Equalization for Single-Carrier Transmission are as follows:

- Enhanced joint channel estimation and signal detection makes the new technique effectively mobile and thus one can easily get network wherever he goes because of continuous handover (due to training).
- An 8 x 8 AMUD MIMO OFDM provides a 2dB SNR gain compared to the conventional MIMO OFDM.
- The sum rate capacity result in the new technique is very close to MIMO theoretical upper bound Fig.5.
- As the modulation order increases, the iterative gain increases.

SYSTEM STRUCTURE FOR AMUD MIMO OFDM

Fig. 1 is the system model for AMUD MIMO OFDM, with Nt and Nr transmit and receive antennas with k subcarriers in one OFDM block. At time t, a data $b'[n,k] : k =$ 0; 1,…, n, transformed into different signals 0,… k-1 and i = 1; 2,…,n, and i are numbers of sub channels of OFDM system. Signals transmitted are modulated by x1[n, k]. The FFT received at each receive antenna is the superposition of the transmitted signals. The receive signal at jth receive antenna is

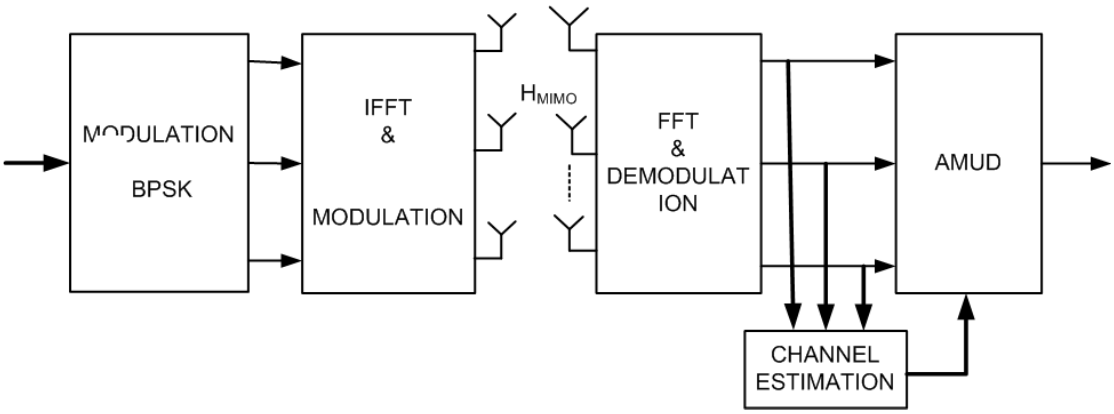


Fig. 1. System Model(AMUD MIMO OFDM)[13]

$$y_j[n,k] = \sum_{i=1}^{N_t} H_{i,j}[n,k] x_i[n,k] + n_j[n,k] \qquad (1)$$

$H_{i,j}[n,k]$

$j,\ n_j[n,k]$

where is the frequency response between antennas i and is the additive Gaussian noise with zero mean and unit variance σ_n^2 .

Turbo Equalizer structure

The received signal r that is sampled at the symbol rate can be given by the equation

$$r = Ah + w \qquad (2)$$

where A is the matrix containing symbols. The channel impulse response is described by the vector $\mathbf{h} = (h0, h1, hL)^T$, which consists of symbol spaced complex valued channel taps.

The white noise samples are denoted by w, the noise variance is $\sigma^2 = N0/2.$

PRINCIPLE OF TURBO QUALIZATION

The iterative equalization receiver structure in Fig. 2, shows that both the equalizer and the decoder employs the optimal symbol by symbol Maximum A-Posterior (MAP) soft input soft output (SISO) algorithm (BAHL 1996). Soft input symbols are fed into the decoder from a sampled receive filter stream r(t) and piece of hard decisions are produced as the final output. It is possible to equalize and decode in an iterative manner that is similar to turbo decoding. The equalizer provides soft outputs, i.e., reliability information on the coded bits for the channel decoder. The soft information on the bit ck is usually given as a log-likelihood ratio (LLR) or L-value

$$\lambda^E(c_k) = log\frac{P((c_k) = +1|r)}{P((c_k) = -1|r)} \qquad (3)$$

which is the ratio between the conditional bit probabilities in the logarithmic domain. These L-values are deinterleaved and given for the channel decoder, which uses them to recover the information bits u. At the first iteration round there is no feedback information from the channel decoder available, so the equalizer calculates the L-values $\lambda^E(c')$ as given by (5) that are just based on the received samples r from the channel. The log-likelihood ratio (LLR) are deinterleaved to break consecutive bits far apart and thus giving the channel decoder independent input values. The interleaving is an essential part in the iterative receiver scheme, since the extra information on an individual data bit is due to the different neighboring bits in the detection and decoding processes of the feedback branch to the equalizer. Therefore we need to use the more complex SISO decoder instead of the conventional hard output decoder. The equalizer is able to produce the L-values $\lambda^E(c')$ based on the received samples from the channel, so that information should not be repeated in the feedback. Hence, the feedback only contains the extra information that is obtained from the surrounding bits in the channel decoding. The input L-values and the obtained extra information are called intrinsic and extrinsic information,

$$\lambda_a^D \bar{c}_k = \lambda^D(\bar{c}) - \lambda_a^E \bar{c}_k \qquad (4)$$

$\lambda_a^E c_k$

respectively. The extrinsic information from the channel decoder is given as (Bauch 1998), (Tuchler 1998).

where denotes the extrinsic information from the equalizer. The turbo equalization technique is based on the utilization of this extrinsic information at the next iteration round (Bauch 1998). So it is passed through the interleaver to the equalizer as a priori information on the bit radiabilities. By exploiting this side information in the detection, more reliable decisions are achieved. Also in the equalizer output the extrinsic information $\lambda_a^E c_k$ is extracted from the output as follows:

$$\lambda_a^E c_k = \lambda^E(c) - \lambda_a^D c_k \qquad (5)$$

This equalizer information is again used in the SISO decoder to produce new soft outputs and furthermore, the new extrinsic information according to (4). As soon as this feedback information becomes available, the new iteration round can be started. The number of iterations may depend on the processing power available or the achieved performance improvement. At the final stage, there is no need for the SISO decoder, since only hard decisions on the information bits are needed. The Turbo equalization receiver is able to improve the performance, but at the cost of higher complexity. The main burden is the complex SISO decoder, especially due to the coding schemes that are based on the constraint length of 4. Also, as the equalization and decoding are performed several times, the receiver complexity grows respectively.

ADAPTIVE MMSE RECEIVER FOR MIMO OFDM

In adaptive filter the parameters are continuously changing due to the received training sequence from the transmitter which informs the receiver to adjust the parameter of the filters to match the desired signal. In the single user environment, every match filter Maximum Likelihood (ML) plays the role of an Adaptive MMSE ML receiver (Rapajic 1994), (Rapajic 1999).

An adaptive turbo receiver structure used in this paper shows that the a bank of adaptive linear MMSE filters trains the filter coefficients and retrieves the signatures. Training is employed using least mean square (LMS) approach to adaptively update the linear coefficient, an MMSE convergence of the filter coefficients provide estimates of the received signatures. An adaptive MMSE filter minimize the error by an adaptive algorithm, the steepest decent algorithm is used to minimize the mean square error (MSE). For simplicity, fractionally spaced adaptive linear transversal filter for Adaptive MMSE detection is used, which is insensitive to the time

$y_3[n,k]$

differences in the signal arrival times of different users, thus the receiver timing recovery is extremely simplified[(Rapajic 1994), (Hana 1991). Consider the received signals $y_1[n,k]$, $y_2[n,k]$ and and let their general form for any node and any path in the network is $y_N[n,k] = r_n[m]$. Where n is the specific number assigned to the signals at nodes. $\mathbf{y_1}[n,k]$, $\mathbf{y_2}[n,k]$ and $\mathbf{y_3}[n,k]$ received digital output symbol block from adaptive filters is $b'[n,k]$.

In multi-cellular environment transmitter transmits information independently. Therefore, non orthogonally transmitted signal from independent users arrives asynchronously at the receiver (Y.Li 2002), (Wittneben 1993) and the delay cannot be neglected. Due to non orthogonality of the spreading code the correlation exist between the spreading code at the receiver. The co-efficient of correlation is given by $\mathcal{E}[\mathbf{r}_M[m]\mathbf{r}_N^*[m]] = \varrho(M,N) = \mathbf{R}_{(M,N)} = \int s_M s_N dt$ and **R** correlation matrix. The digital output of the nth filter for the mth symbol period on relay or destination is given by

$$\mathbf{r}_n[m] = \mathbf{R}\mathbf{H}\mathbf{x}_n[m] + \mathbf{v}_n[m] \tag{6}$$

$$\epsilon = (\mathbf{x}_n[m] - \hat{\mathbf{x}}_n[m])$$

$$\epsilon = ([\mathbf{x}_n[m] - \mathbf{a}_n^H[m]\mathbf{r}_n[m]) \tag{7}$$

$\mathbf{a}_n^H[m]$

where **H** is matrix of respective channels and vn[m] is noise. The error between the reference signal and the output of adaptive filter is is M dimensional complex valued weight vector at mth symbol time when the variable filter estimates the desired signal $a_n[m] = [a_1, a_2, \ldots\ldots, a_M]$ by convolving the input $r_n[m] = [r_1, r_2, \ldots\ldots, r_M]$. signal with impulse response. M are tap of filter and During the adaptation mode

$$J_{a_n} = \mathcal{E}[\epsilon_n \epsilon_n^*] \tag{8}$$

from equation (11)

$$J_{e_n} = \mathcal{E}[(\mathbf{x}_n - \mathbf{a}_n^H\mathbf{r}_n)][(\mathbf{x}_n - \mathbf{a}_n^H\mathbf{r}_n)^*] \tag{9}$$

$$J_{e_n} = \mathcal{E}[\mathbf{x}_n\mathbf{x}_n^*] + \mathbf{a}_n^H\mathcal{E}[\mathbf{r}_n\mathbf{r}_n^H]\mathbf{a}_n - \mathbf{a}_n^H\mathcal{E}[\mathbf{r}_n\mathbf{x}_n^*] - \mathcal{E}[\mathbf{x}_n\mathbf{r}_n^H]\mathbf{a}_n \tag{10}$$

$\mathcal{E}[\mathbf{x}_n\mathbf{x}_n^*]$,

the weight parameters are adjusted such that mean square error J_{a_n} is minimized in mth symbol time. For simplicity of description m is with every term but we are not mentioning it here $\mathcal{E}[\mathbf{r}_n\mathbf{r}_n^H]$The first term in equation represents the variance of desired signal. The expectation denotes M by N correlation matrix of the received signal given by

$$R = \begin{pmatrix} \varrho(1,1) & \varrho(1,2) & \ldots & \varrho(1,N) \\ \varrho(2,1) & \varrho(2,2) & \ldots & \varrho(2,N) \\ \ldots & \ldots & \ldots & \ldots \\ \varrho(M,1) & \varrho(M,2) & \ldots & \varrho(M,N) \end{pmatrix} \tag{11}$$

The matrix **R** is Hermitian and can be uniquely defined by specifying the values of the correlation coefficients where

$$\mathcal{E}[\mathbf{x}_{(n)}\mathbf{r}^{H}_{(n)}] = Z^{H}.$$

$$Z^{H} = [z_1^*, z_2^*, \ldots\ldots z_M^*].$$

$$z_i = \mathcal{E}[\mathbf{r}_{(n)}\mathbf{x}^*_{(n)}].$$

The expectation is M by N cross-correlation matrix vector between the received components and the reference sequence, and expectations where The co-efficient is given by

$J_{(a_n)}$

$\nabla_{n}.$

$$\nabla_n = \frac{\partial J(a_n)}{\partial a_n} = \begin{pmatrix} \frac{\partial J(a_n)}{\partial a_1} \\ \frac{\partial J(a_n)}{\partial a_2} \\ \cdots\cdots \\ \frac{\partial J(a_n)}{\partial a_M} \end{pmatrix} \quad (13)$$

For stationary input and reference signals the surface obtained by plotting the mean square error $J_{(a_n)}$ versus the weight co-efficient has a fixed shape and curvature ith a unique minimum point; the adaptive process seeks that minimum point at which the weight vector is optimal. Differentiating the mean squared error function with respect to each coefficient of the weight vector an yields the gradient

$\nabla_{n}.$

$$\nabla_n = -2Z + 2Ra_n = 0 \quad (14)$$

The optimal weight vector aopt can be determined by setting the gradient equal to zero, where 0 is an M by 1 null vector at the minimum point of the error surface, the adaptive MUD is optimum in the mean squared error sense, and equation can be simplified in the form $Ra_{opt} = Z$ Which is Weinner-Hopf equation or the normal equation, where the vector representing the estimation error is normal to the vector representing the output of the combiner. One possible solution of this equation is matrix inversion

$$Z = R^{-1}a_{opt} \quad (15)$$

$$a_n = a_n - \mu\nabla_n \quad (16)$$

μ

Another simple solution that does not require matrix inversion or explicit calculations of the correlation coefficients is the steepest decent method. The Steepest Decent Method is recursive procedure that can be used to calculate the optimal weight vector a_{opt}. Let an and rn denote the values of the weight vector and the gradient vector at time m, respectively. Then succeeding values of the weight vector are obtained by the recursive relation. After each symbol period m the weight of the filter updated till optimum coefficient get best cross correlation value. The filter can go to decision directed mode where coefficient continuously change with the variation of channel

Where is step size constant that controls stability and the rate of adaptation. If we express rn in terms of instantaneous estimates $Z = r_n x_n^*$ and $R = r_n r_n^H$.

$$a_n[m+1] = a_n[m] + 2\mu r_n[m][x_n^*[m] - r_n^H[m]a_n[m]] \quad (17)$$

$$\epsilon_n^*[m]$$

$$a_n[m+1] = a_n[m] + 2\mu r_n[m]\epsilon_n^*[m] \quad (18)$$

Then the equation can be simplified as which can be expressed in term of as
Here is correction factor, where m= achieved.
The equation
$2\mu r_n[m]c_n^*[m]$

$0, 1, 2.......$ till a_{opt}

explains that the updated weight vector is computed from the current weight vector by adding the input vector scaled by the complex conjugate value of the error and by μ which controls the size of correction. The iteration of the equation produces the value of the Mean Square Error at which the vector tends to its optimal value a_{opt}. Minimum error value cannot be reached by a finite number of iterations though approachable. To achieve proper adaptation, the weight vector must be updated at a rate fast enough to track the channel variations. The method of steepest descent can be viewed as feedback model which may become unstable. The stability of the steepest descent algorithm depends on the step size parameter μ and the auto correlation matrix R. The eigenvalues of R are all real and positive, the condition for convergence and stability of the steepest descent algorithm depends on the step size parameter μ.

ADAPTIVE CAPACITY EQUATION

System capacity will be significantly improved by MIMO channels (Teletar 1999), (Winters 1987). In OFDM (Cimini 1985), (Zelt 2000), (Y.Li 1999), the

$$C = log\left(\frac{1}{\sigma_e^2}\right) \quad (19)$$

σ_e^2

entire channel is divided into many narrow parallel sub channels. The capacity formula for the proposed scheme as stated in (Rapajic 1999), based on the assumption that the channel matrix which consists of independent and identically distributed iid Rayleigh fading coefficients and Fig.4 is the sum rate capacity figure of the developed scheme which is very close to MIMO capacity. The capacity formulae for adaptive multiuser detection by Predrag is as follows:
where is the noise variance of the signal at the receiver.

PERFORMANCE COMPARISON AND SIMULATIONS

AMUD OFDM MIMO provides a fair comparison with some parameters held at a constant during simulation (pilotbits, spreading factor and frame size). An achieved bit error rate at 10□5 in the following three cases were compared by Monte Carlo simulation, with assumed perfect channel state information, BPSK Modulation and flat fading channel model were employed in the simulation. The configurations considered for OFDM system with 64 subcarriers, 16 symbol time periods and 4 symbol period for antenna configuration Nt = Nr, as shown in antenna configurations 2x2 and 8x8.

Fig3. and 4. Shows the SNR in dB versus the BER of the SISO OFDM, the 2 X 2 MIMO and AMUD OFDM at a significant SNR gain, while the 8 x 8 MIMO and AMUD MIMO OFDM provides a 2dB gain indicating the higher the number of the antenna the better the performance. AMUD OFDM MIMO performs better than other schemes and because of it's continues handover in comparison to the conventional MIMO OFDM scheme. Fig.5, is the sum rate capacity comparative results of the proposed technique with the other schemes in bits/second/Hz which is very close to MIMO capacity upper bound.

In the BER performance of the iterative equalizer, it depends upon the channel profile, unlike that of the AMUD, the modulation scheme, the encoder constraint length and the size of the interleaver. In this paper the encoder constraint length and the size of the interleaver are fixed. For channel coding, we used a rate 1/2 recursive systematic convolution code with memory mc = 2, constraint length L = 4, generator polynomial G = [75], Block length = 4096. Fig. 6 and 7, shows the performance of the BER as a function of signal to noise ratio (SNR) for BPSK, QPSK, 8PSK and 16QAM modulation schemes for memory two and memory 4 respectively, using channel model B and decoded with MAP algorithm (Bauch 1995)

These performance curve are important, as they represent an upper bound performance for the iterative equalization receiver. When the respective bound are met, this indicate complete mitigation of inter symbol interference (ISI) and Multiuser Interference (MUI) introduced by the channel. If we take a target of 10^{-4}, then to achieve this target BPSK mode required 5dB, QPSK 5.5dB, 8PSK 7.5dB, 16QAM 8.4dB and in memory 4 BPSK mode is 4dB, QPSK 4.7dB, 8PSK 6.3dB, 16QAM 7.6dB respectively. As the modulation order increases, the iterative gain increases.

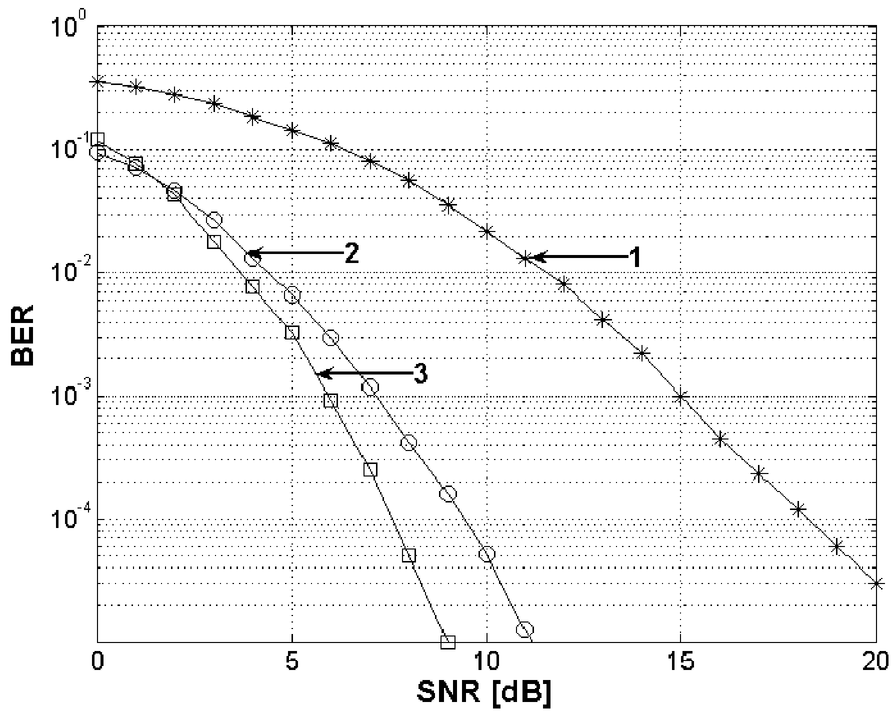


Fig. 3. The schemes 2 × 2 Bit error rate comparison

1) OFDM SISO
2) OFDM MIMO
3) AMUD OFDM MIMO

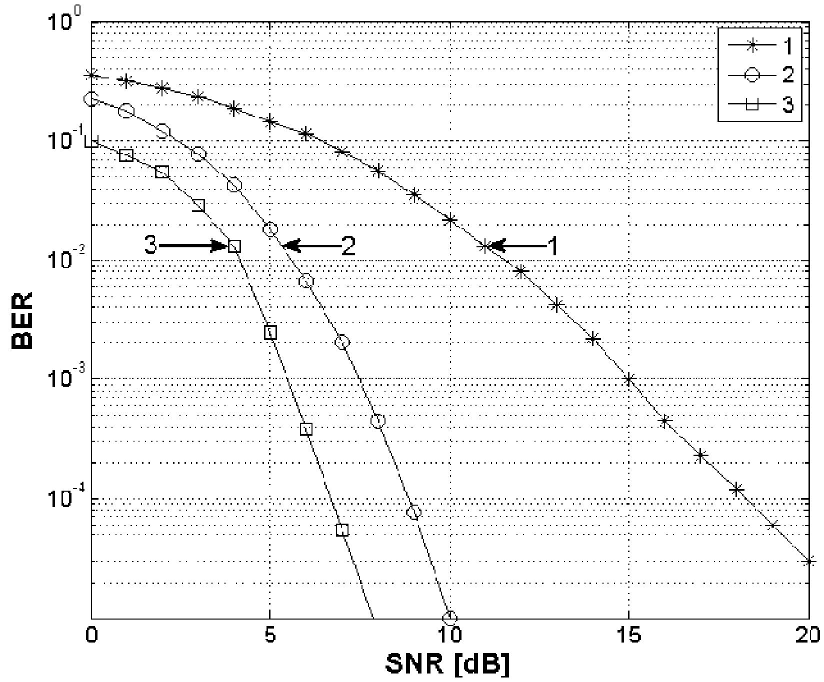


Fig. 4. The schemes 8 × 8 Bit error rate comparison

1) OFDM SISO
2) OFDM MIMO
3) AMUD OFDM MIMO

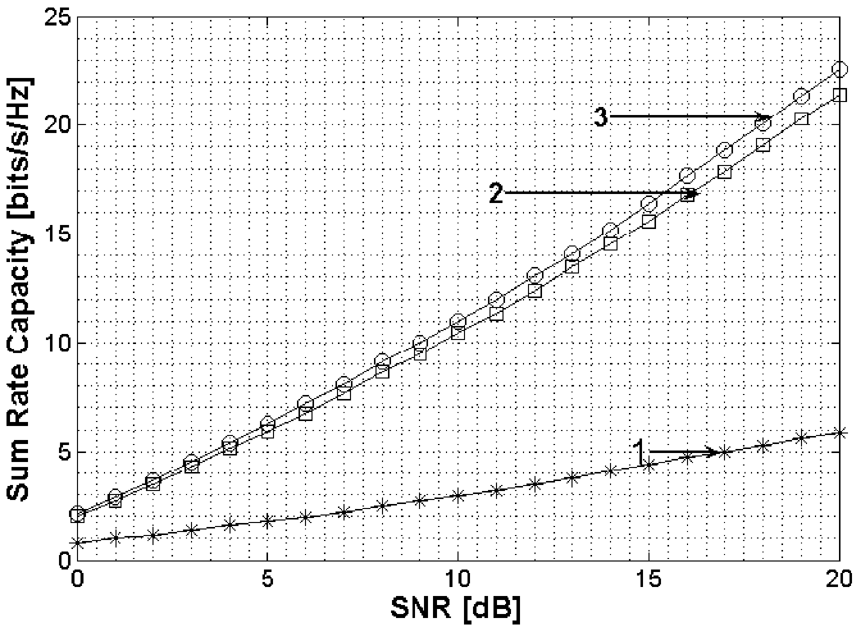


Fig. 5. Capacity comparisons (in bits/sec/Hz) (4 × 4)

1) OFDM SISO-Sum rate capacity
2) OFDM MIMO-Sum rate capacity
3) AMUD OFDM MIMO-Sum rate capacity

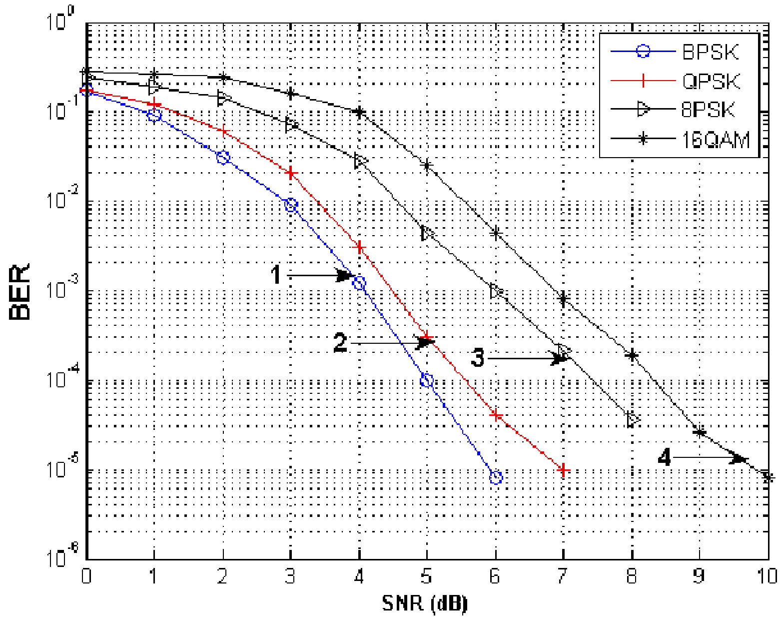


Fig. 6. The schemes memory 2 Bit error rate comparison

1) BPSK
2) QPSK
3) 8PSK
4) 16QAM

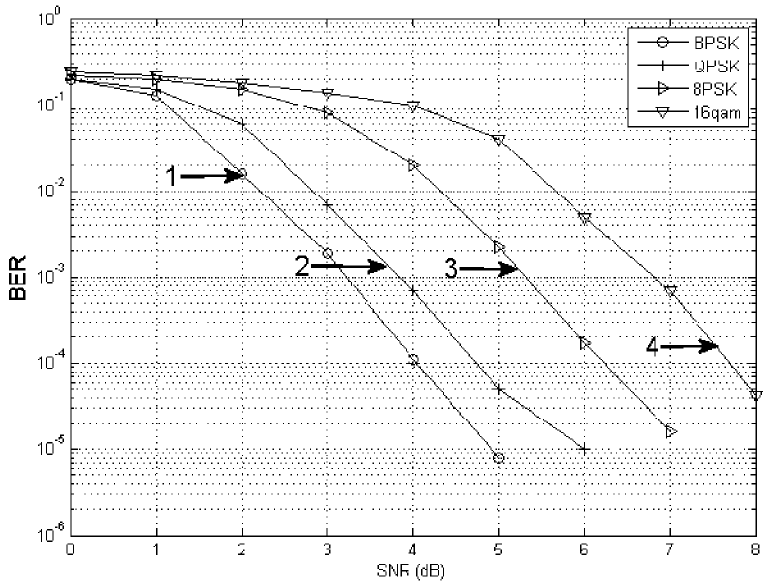


Fig. 7. The schemes memory 4 Bit error rate comparison

1) BPSK
2) QPSK
3) 8PSK
4) 16QAM

CONCLUSION

AMUD MIMO OFDM applies joint detection by implicit assumption of the use of an optimum MUD in comparison to conventional individual parameter estimation by Telatar. The developed technique has good performance in terms of bit error rate, SNR and capacity in Raleigh channel. The schemes 8 x 8 antenna configurations at BER of 105 provide a 2dB SNR gain compared to the conventional MIMO OFDM. The sum rate capacity is very close to MIMO theoretical upper bound (21.5 bits/s/Hz at signal to noise ratio of 20dB).

The results for higher modulation schemes show that as the modulation order increases, a higher SNR is required to obtain the same BER performance at a lower order. They also demonstrate that the iterative gain is greater for higher modulation orders. However, there is a trade off, between the iterative gain for higher modulation orders and the complexity of the receiver. The complexity at the receiver is dominated by the complexity of the MAP equaliser.

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