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Research Article

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Adaptive Beam Forming of MIMO System using Low Complex Selection of Steering Vector

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Abstract

With the superior growth of digital communication the need for high-speed data transmission has increased. In that, Orthogonal Frequency Division Multiplexing (OFDM) is multi-carrier modulation techniques, which partition the single spectrum into numerous sub-carriers. The major benefit of OFDM is their significance towards channel fading in wireless environment. In wide range data transmission, the OFDM is heavily affected by the Inter Symbol Interface (ISI). To diminish ISI, an Adaptive Analog Beam Forming (AABF) based Phased Array Antenna (PAA) is introduced. PAA achieves the reduction of noise significantly and also shows improvement in the effectiveness of reducing the side lobe ranges with narrow beam width. The experimental outcome proves that the proposed system performs effectively than the other existing systems.

Keywords: Adaptive analog beam forming; Orthogonal frequency division multiplexing; Phased array antenna; Channel state information; Inter symbol interface

Introduction

In the past few decades, there has been intensive work in developing effective multiuser transmission methods for wireless system [1]. Multi Users-OFDM (MU-OFDM) has become an emerging technique in fourth and fifth generation wireless communication [2,3]. The MU technology transmits numerous signals over several antennas and OFDM partition the single radio signal into equally spaced sub-signals to deliver less distortion rate at high speed broadcast [4]. The major issue in wireless communication is ISI, several techniques are implemented to decrease ISI [5]. In that, Beam Forming (BF) can effectively improve the reliability of transmission and also achieve high data rate [6]. Generally, BF is performed on the basis of analog and digital. In this research work, the analog BF is preferred for implementation as it improves the performance of target orientation. Initially, BF was implemented in unipolar and bipolar antennas to increase the signal capacity and also for improving the signal strength.

Still, BF requires Channel State Information (CSI), because it helps to denote the channel property of communication link and also to avoid complex equalization at the receiver. Hence, the Rayleigh channel is preferred for CSI. Practically, suitable CSI is not obtainable at the transmitter system and the impact of quantized or noisy CSI is very excessive [7]. To enhance the spatial diversity gain or to reduce the cross correlation of the signals reflected back to the antenna by the target of interest, a flexible system was designed named Phased Array Antenna (PAA) [8]. It has the capacity to direct the beam pattern electronically with extraordinary effectiveness and also it manages to get minimal side lobe level with narrow beam width [9]. To achieve scanning range with high angle resolution, an enormous number of antenna components are required to implement the array. Usage of the microwave components in an enormous quantity may cause frequent obstacles in limited coverage areas, and also it gets complex during deformation. So, to overcome these concerns a traditional PAA is improved as adaptive PAA, which enhance the antenna gain and improves the directivity of the main lobe [10].

Research Method

This paper evaluates the directivity of the antenna that is enhanced by adaptive PAA by reducing the side lobes and also by decreasing the antenna components. Generally, the radio signals are spread out in all directions by a distinct antenna and similarly a distinct antenna will collect signals evenly from all directions. Assume, a linear antenna array sensors and narrowband signals received at snapshot as expressed by:

$$x(i)=V(\theta)s(i)+n(i). \tag{1}$$

Where, $\theta=[\theta_1,\theta_2.....\theta_k]^T \subseteq R^k$ determines the vector containing the directions of arrival (DoAs) and $s(i)\in C^{K\times 1}$ denotes the uncorrelated source signal. The $V(\theta)$ represents the matrix containing Steering Vector (SV), which is denoted as $V(\theta)=[v(\theta_1)+e.....v(\theta_K)]\subseteq C^{M\times K}$ and e is the SV for the mismatched signals. Here, $n(i)$ is specified as zero mean and variance of Gaussian noise. Hence, AABF can point beam in many direction and manipulate beam shape to enhance system performance by varying the vector $V(\theta)$.

Analog beam forming

It is expensive to insert a complete RF receiver chain and high-rate high-resolution analog channels unit for each antenna. Thus, let consider an Analog Pre-processing Network (APN) inserted in RF domain immediately after the low-noise amplifiers and bandpass filter. The effect of APN on the baseband signal is modelled using a discrete time equivalent matrix operation. Consider a linear BF signal with the output given by:

$$y(i)=w^H x(i) \tag{2}$$

Where, $w=[w_1....w_M]^T \in C^{K\times 1}$ is denoted as beam-former weight vector and $(\bullet)^H$ indicates the Hermitian Transpose. Optimum beam-former can be computed by maximizing the SINR, which is given by,

$$SINR=\frac{\sigma_1^2\left|w^H a\right|^2}{w^H R_{i+n} w} \tag{3}$$

Where, summarized SV is stated as a and the Desired Signal Power (DSP) is mentioned as σ_1^2, INC matrix is mentioned as R_{i+n}. Then the

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issue in the equation (3) can be converted into an optimization issue, which is generally named as MVDR beam-former or Capon beam-former.

$$\underset{w}{\text{minimize}}\ w^H R_{i+n} w,$$

$$\text{subject to } w^H a = 1 \quad (4)$$

The INC matrix R_{i+n} is found by determining the Sample Covariance Matrix (SCM) of the received data as,

$$\hat{R}(i) = \frac{1}{i}\sum_{k=1}^{i} x(k)x^H(k) \quad (5)$$

Implementation of phased array antenna (PAA)

The PAA is composed by similar group antennas, each antenna consists of summing network, phase shifter, power feed network and that helps to represent a beam on a desired location. In RF stage, the analog components like low noise and power amplifiers are required, to condition the both transmitted and received signals in antenna array. Incident plane wave of antenna receiver can be demonstrated by following equation,

$$f(t,p_n) = C_n(t) = x(t-\tau_n)\cos(w_{RF}(t-\tau_n)), n = 0,.....N-1 \quad (6)$$

$$\approx x(t)\cos(w_{RF}t - \theta_n) \quad (7)$$

Where, θ_n is given by,

$$\theta_n = w_{RF}\tau \quad (8)$$

After that, the respective incoming signals reaches RF modulation stage. Here, the frequency components of incoming signals are associated to ADCs and analog signal modulation, is needed to shift the signal's frequency components into a lower frequency band. If the RF stage has local oscillator with a frequency of w_{LO}, then the modulated beam from PAA can be termed in the following form,

$$g_n(t) = x(t)\cos(w_{RF}t - \theta_n)\cos(w_{LO}(t)) \quad (9)$$

Where, k is modulated as time signal t, analog beam $y(i)$ is replaced by the PAA's analog beam $g_n(t)$

Estimation of steering vector (SV) using proposed PAA analog beam

The cross-correlation relationship between the array observation vector and the beam-former output can be specified as follows,

$$d = E\{xg_n\} \quad (10)$$

Where, the vector v is considered as $|v_m^H w| << |v_1^H w|$ for $m = 2,.......K.$ Hence, all the source and noise signals have zero mean, substitute the equation (1) and (9) in equation (10), which is mathematically stated as,

$$d = E\{\sigma_1^2 v_1^H w v_1 + tw\} \quad (11)$$

In which the unwanted interference that cause noise and signal power loss are eliminated. Then the SV is projected into the sub-space and the projection matrix P is calculated as,

$$P = [c_1, c_2....c_p][c_1, c_2....c_p]^H \quad (12)$$

Where, $c_1, c_2....c_p$ are the P Eigen vectors from the vector of matrix C, which is given as,

$$C = \int_{\theta_1-\theta_e}^{\theta_1+\theta_e} v(\theta)v^H(\theta)d\theta \quad (13)$$

Oracle Approximating Shrinkage (OAS) technique is utilized by the PAA method, to evaluate the correlation vector. An accurate estimation helps to achieve a better estimate of SV. It can be specified as,

$$\hat{F} = \hat{r}I \quad (14)$$

Where, $\hat{r} = tr(\hat{S})/S$ and $\hat{S} = diag(xg_n)$. A reasonable trade-off between reduction in covariance and bias growth can be attained by shrinkage of $\hat{S}$ and $\hat{F}$, results in

$$\hat{d} = \hat{\rho}diag(\hat{F}) + (1-\hat{\rho})diag(\hat{S}) \quad (15)$$

Which is parameterized by the shrinkage co-efficient $\hat{\rho}$, utilized for reducing the Mean Square Error (MSE) and the Sample Correlation Vector (SCV) is specified in the equation (16) as,

$$\hat{S}(i) = diag\left(\frac{1}{i}\sum_{k=1}^{i} x(k)g_n(k)\right) \quad (16)$$

Once the correlation vector $\hat{d}$ is attained by the above OAS technique, the SV is estimated by,

$$\hat{a}_1(i) = \frac{P\hat{d}(i)}{\left\|P\hat{d}(i)\right\|_2} \quad (17)$$

Where, gives the final estimate of SV and by identifying the SV, beam formed by the PAA is estimated. Thus the PAA implementation and the SV are identified for the analog signals and the adaptive analog beam-former is detected.

$$\hat{d}(i) = \hat{p}(i)diag(\hat{F}(i)) + (1-\hat{p}(i)diag(\hat{s}(i)), \quad (18)$$

$$\hat{p}(i+1) = \frac{(1-\frac{2}{M})tr(\hat{D}(i)\hat{S}^*(i)) + tr(\hat{D}(i))tr(\hat{D}^*(i))}{(i+1-\frac{2}{M})tr(\hat{D}(i)\hat{S}^*(i)) + (1-\frac{i}{M})tr(\hat{D}(i))tr(\hat{D}^*(i))}, \quad (19)$$

Interference plus-noise covariance matrix (IP-NCM) determination

To evaluate INC matrix, the data covariance matrix is required and the SCM in equation (5) is essential as an initial approximation. In the next step, similar to OAS estimate, the cross-correlation vector d is performed with OAS technique for further shrinkage determination step. It is defined as,

$$\hat{F}_0 = \hat{v}_0 I \quad (20)$$

Where, $\hat{v}_0 = tr\hat{R}/M$ and utilize the shrinkage form again,

$$\hat{R} = \hat{p}_0\hat{F}_0 + (1-\hat{p}_0)\hat{R} \quad (21)$$

By reducing the MSE, which is termed as $E\left\|\hat{R}(i) - \hat{F}_0(i-1)\right\|^2$, then obtain the following recursion,

$$\hat{R}(i) = \hat{p}_0(i)\hat{F}_0(i) + (1-\hat{p}_0(i)\hat{R}(i) \quad (22)$$

$$\hat{p}_0(i+1) = \frac{(1-\frac{2}{M})tr(\hat{R}(i)\hat{R}(i)) + tr^2(\hat{R}(i))}{(i+1-\frac{2}{M})tr(\hat{R}(i)\hat{R}(i)) + (1-\frac{i}{M})tr^2(\hat{R}(i))}, \quad (23)$$

in (22) and (23) is guaranteed to converge (18). To eradicate the unwanted information of desired signal in the covariance and INC matrix, the DSP σ_1^2 must be estimated. Received data can be modified as,

$$x = \sum_{k=1}^{k} a(k)s(k) + n \quad (24)$$

Pre-multiplying the equation (24) by a_1^H, which is illustrated as,

$$a_1^H x = a_1^H a_1 s_1 + a_1^H(\sum_{k=1}^{k} a(k)s(k) + n) \quad (25)$$

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Considering that a_1 is un-correlated with the interferers, equation (25) is updated as,

$$a_1^H x = a_1^H a_1 s_1 + a_1^H n \quad (26)$$

Assuming the expectation of $E\left\|(a_1^H x)^2\right\|$, where equation (26) is expanded as follows,

$$({}_1^H x)^2 = E[(a_1^H a_1 s_1 + a_1^H n)*(a_1^H a_1 s_1 + a_1^H \quad (27)$$

If the noise is independent of the desired signal and it is mathematically stated as,

$$(a_1^H x)^2 = (a_1^H a_1)^2 (s_1)^2 + a_1^H n n^H a_1, \quad (28)$$

Where, $(s_1)^2$ is the DSP, which is inter-changed by its estimate $\hat{\sigma}_1^2$, and $n n^H$ is represented as NCM R_n, which is inter-changed by $\hat{\sigma}_{1IM}^2$. Replacing, the general estimate a_1 as DSP estimate $\hat{a}_1(i)$, which is demonstrated as,

$$\lambda = \frac{2(\hat{\sigma}_1^2(i)\hat{a}_1^H(i)\hat{a}_1(i) - g(i)x^H(i)\hat{a}_1(i))}{\hat{a}_1^H(i)\hat{a}_1(i)} \quad (29)$$

Finally, the desired signal of covariance matrix is subtracted and the INC matrix is stated by

$$\hat{R}_{i+n}(i) = \lambda \hat{R}(i) - \hat{\sigma}_1^2(i)\hat{a}_1(i)\hat{a}_1^H(i). \quad (30)$$

Where, λ is deliberated as a weight function,

In AABF scheme, a weight function is employed for minimizing the BER. Usually, BER is inversely associated with the quality of signals. Once the BER is minimized means, the quality of the signals gets improved.

$$\hat{\sigma}_1^2 = \frac{\left|\hat{a}_1^H(i)x_1(i)\right|^2 - \hat{a}_1^H(i)\hat{a}_1(i)\hat{\sigma}_n^2}{\left|\hat{a}_1^H(i)a_1(i)\right|^2} \quad (31)$$

wing approaches, it does not need direction finding. With the estimates for the SV and the INC matrix, the beam-former is computed by

$$\hat{w}(i) = \frac{\hat{R}_{i+n}^{-1}(i)\hat{a}_1(i)}{\hat{a}_1^H(i)\hat{R}_{i+n}^{-1}(i)\hat{a}_1(i)}, \quad (32)$$

Results and Discussion

Simulation and analysis of the outcome is done by employing Uniform Linear Array (ULA) $M = 12$, with a space of half wavelength. Initially, three source signals are added to the desired signal, the first one is recognized at $\theta_1 = 10^\circ$ and the remaining two interferers are interrupting on the antenna array from the directions $\theta_2 = 50^\circ$ and $\theta_3 = 90^\circ$. The Signal-to-Interference Ratio (SIR) is fixed at 20 dB. Only one iteration is achieved per snapshot and employs $i = 50$ snapshots and 100 repetitions to obtain each point of the curves.

The beam-former computed with Low-Complexity Shrinkage-Based Mismatch Estimation Algorithm (LOCSME) is compared to existing beam-formers in terms of output SINR. For the beam-formers of LOCSME, inverse LOCSME, and the proposed scheme of AABF on Phased antenna, the angular section is chosen as $[\theta_1 - 5^\circ, \theta_2 + 5^\circ]$ and $p = 8$ principal eigenvectors are used. Eigenvectors of the subspace projection matrix P is selected manually with the help of simulation.

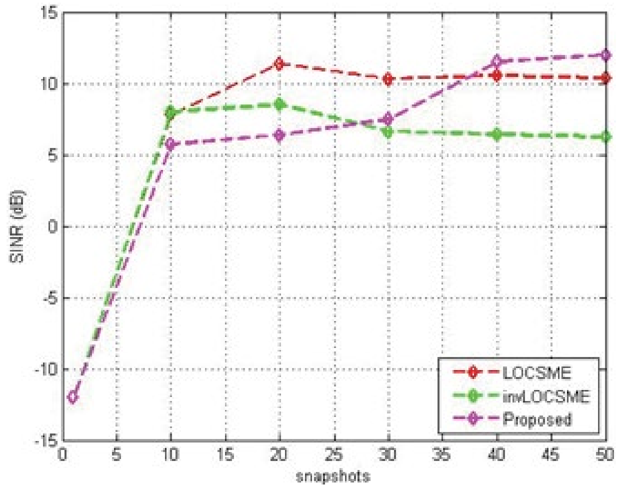

Figure 1: SINR vs. snapshot.

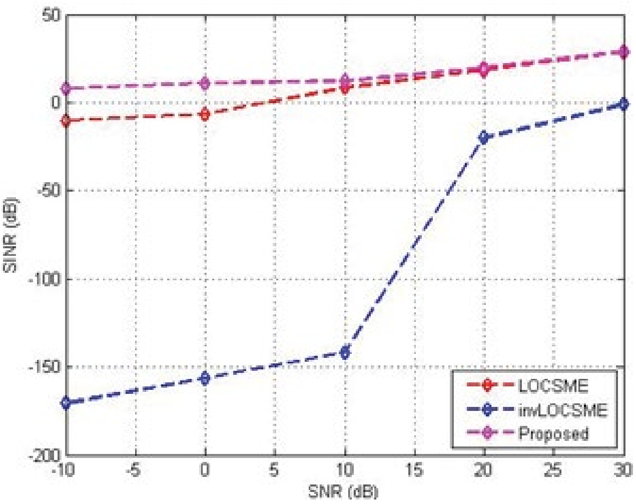

Figure 2: SINR vs. SNR.

For the beam-formers of LOCSME, inverse LOCSME and the proposed approach, which requires an optimization technique, the matlab CVX software is employed. Algorithms (SINR vs snapshot and SINR vs SNR) performance are shown in Figures 1 and 2. In Figure 1, the existing schemes LOCSME and inverse LOCSME shows 11dB, and 6 dB respectively. Compared to these two schemes, the proposed performance proves with a significant outcome of 12dB. The AABF technique confirms that the number of snapshots increases means the SINR value is also improved.

In Figure 2, SINR performance of the existing two methods and the proposed system (AABF) is signified in a graphical form. The number of snapshot is 50 for (SINR vs SNR) plots and the regular execution time of the algorithms in the early methods are around 0.3 sec/snapshot, where the proposed scheme requires only 0.016 sec/snapshot.

In Figure 3, BER performance of the existing two methods and the proposed approach is denoted in a graphical form. It specifies that the performance of proposed approach (AABF is -11dB) is superior in terms of SNR.

$$\Sigma NP(\delta B) = 10\log\left(\frac{\hat{p}_{signal}}{\hat{p}_{noise}}\right) \quad (33)$$

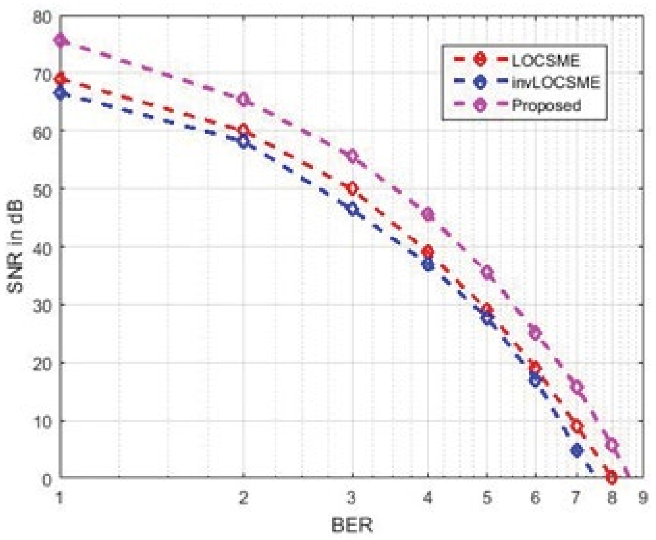

Figure 3: BER vs. SNR.

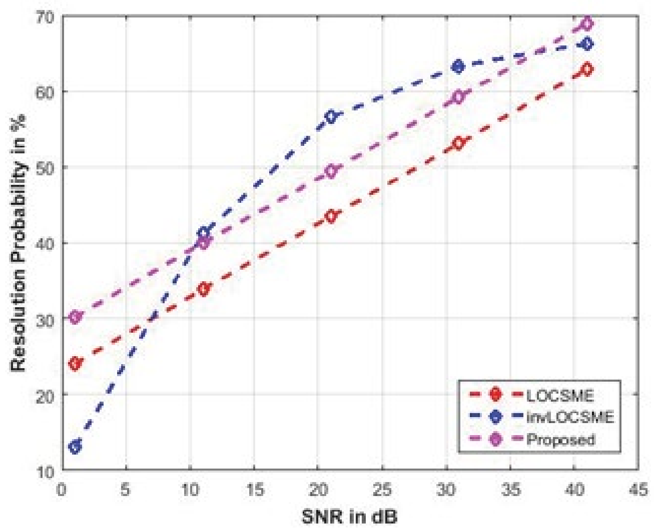

Figure 4: Resolution probability vs. SNR.

Usually, OFDM converts the multi-path channel into N fading channel. Then, derive the BER equation over Rayleigh channels $Pb(E)$. General equation for BER is specified as,

$$\text{BEP} = / N\sum_{K=1}^{N} Pb, K(E) \qquad (34)$$

Where, K is demonstrated as sub-carrier index. Following graph determines the cumulative outcome, because it varies based on the number of snapshots utilized.

Mismatch due to coherent local scattering (CLS)

Figures 1 and 2 illustrate (SINR vs snapshot and SINR vs SNR) performance under CLS case. The proposed AABF method out-performs the other algorithms, which is near to the optimal SINR.

Mismatch due to incoherent local scattering (ILS)

Figures 1and 2 depicts (SINR vs snapshot and SINR vs SNR) performance. Compared to CLS, all the algorithms have performance reduction due to ILS effect. However, the proposed methodology is able to out-perform the remaining robust beam-formers over an extensive range of input SNR. The reason for the improved performance of proposed scheme is the combination of INC matrix and SV mismatch in array antenna. Further, testing with a higher number of antenna array elements specifies that the performance of all algorithms degrades (e.g. AABF in PAA) has around 2 dB degradation when $M = 60$. In addition, inappropriate choice of angular sector also leads to noticeable performance degradation.

Figure 4 specifies the enactment evaluation of proposed and existing approaches in terms of resolution probability. It clearly shows that the proposed scheme delivers 69.90% of resolution probability, which is 4.40% and 7.90% more than the existing approaches inverse-LOCSME and LOCSME. From the Figure 4, the proposed method and LOCSME shows linear in variation, but the inverse-LOCSME scheme delivers random variation that is due to random sequence generation. Here, the number of sensor undertaken is 12 and the number of interference is 10.

Figure 5 determines the antenna gain in terms of sensors, number of sensors undertaken 12. In this work, number of interference and snapshots utilized for experimental analysis are 3 and 50 respectively. Figure 5, clearly shows that the proposed scheme outperforms existing approaches. Whereas, the proposed method gives 5.41 dB antenna gain

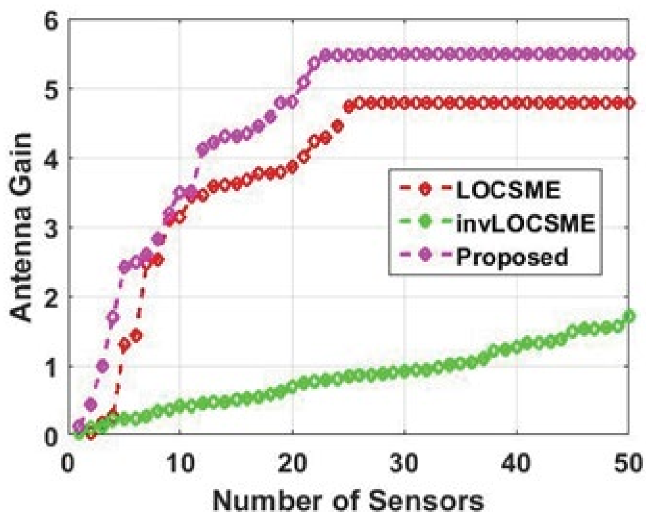

Figure 5: Antenna gain vs. number of sensors.

and the existing schemes shows 4.87 dB and 1.76 dB gain. Hence, the peak gain for dipole antenna is 7.85 dB. This proposed method almost achieved a better antenna gain only with a difference of 2.44 dB in comparison with ideal antenna.

Conclusion

This paper presents a simple, effective adaptive BF algorithm, which is robust against covariance matrix uncertainty. In MU-OFDM, the concern of ISI is continually affecting the signal transmission performance by creating numerous noise, distortion and variations in the BF signals. Here, (SINR vs snapshot and SINR vs SNR) performance is evaluated and shown as a graphical representation. Simulation result confirms that the proposed PAA out-performs in an efficient manner in the both ranges of CLS and ILS cases. Further testing with a higher number of antenna array elements specifies that the performance of proposed PAA is considerably significant than the previous approaches.

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