Design and Performance Analysis of Smart Antenna System for DECT Radio Base Station in Wireless Local Loop

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Abstract—**Digital Enhanced Cordless Telecommunication (DECT) can be a latent solution for wireless local loop (WLL) based communication system planning. However DECT performs poorly in multipath propagation scenarios. To overcome the difficulties and utilize the advantages of DECT systems, smart antennas can be introduced. In this paper, the design and simulation of a smart antenna system for DECT radio base stations in WLL is presented. A 8×8 planar microstrip antenna array is designed and associated signal processing techniques for one and two dimensional cases of smart antenna systems are analyzed. Two different algorithms are used for Direction-of-arrival (DOA) estimation using the 8×8 array parameters, these are MUltiple SIgnal Classification (MUSIC) and Estimation of Signal Parameters via Rotational Invariance Techniques (ESPRIT). Simulation results and performance comparison of these two algorithms for optimum output are incorporated. For adaptive beamforming, Least Mean Square (LMS) algorithm is used here. Radiation characteristics, gain and return loss of the fixed beam planar array antenna are simulated using Zeland IE3D software. Signal processing simulations are run in MATLAB. This smart antenna system is designed for DECT system in 1.88−1.90GHz frequency band.**

Index Terms—**DECT, LMS, MUSIC, WLL, Planar array, Smart antenna System**.

I. Introduction

To exploit the dramatic advent of information and communication technology (ICT) into distant areas Wireless Local Loops (WLL) plays a vital role besides mobile communications. In contrast to Plain Old Telephone System (POTS) or broadband internet connection where copper wires are used to connect end users to backbone network, WLL uses wireless link as the last mile solution. Consequently it can provide very cost effective and rapid deployment over vast area which is particularly essential to extend the ICT sector in developing countries like Bangladesh, India, Kenya etc. Using advanced digital radio technologies, WLL can provide a variety of data services and multimedia services as well as voice. DECT system can also be used for outdoor applications; it may be university campus established within large areas, industrial areas in remote places, oil/gas field or suburb/rural area. DECT is also considered WLL, when a public network operator provides wireless service directly to the user via this technology. WLL has two standards – mobile and Fixed or local area network. Several technologies are used for WLL in fixed or local area network like DECT (local loop), LMDS, IEEE 802.11, WiMAX or 802.16 etc. DECT has the advantage of low cost equipment at both user and service provider end. Moreover, it doesn't require the valuable cellular spectrum as operates in 1880-1900 MHz band. DECT is recognized by the ITU as fulfilling the IMT-2000 requirements and thus qualifies as a 3G system [1]. It is also one of the leading cordless technology [2].So for rapid deployment of PSTNs (public switched telephone networks) DECT can be a promising candidate. A simplified model of DECT used as WLL described in reference [3]. The problem is that DECT requires Line of Sight (LOS) communication which might not be as feasible in urban areas as in suburb/rural areas. To significantly improve the communication link in multipath scenario Smart Antenna System can be deployed. Smart antennas have the potential to provide enhanced range and reduced infrastructure costs in early deployments, enhanced link performance as the system is built-out, and increased

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long−term system capacity [4]. Smart antenna system consists of an array of radiating elements able to steer the main lobe beam towards the desired signal and to locate suitable nulls of the radiation pattern in the direction of interferences. Many works have been done on smart antenna system references [5] − [9] are some of these. In reference [5] a complete Smart antenna system was presented for Mobile Ad-Hoc Networks (MANETs). References [6]−[9] describes DOA algorithms and smart antennas for uniform linear array (ULA) and considering signal arrivals in one dimension only. Also the numbers of antenna elements/ sensors are small. Contrasting to most of the earlier published work that cover only one dimensional conditions for DOA estimations and smart antennas, this paper presents an inclusive effort on smart antennas that incorporate planar antenna array design, the development of signal processing algorithms for angle of arrival estimation (both azimuth and elevation angle) and adaptive beam forming techniques. We use two different algorithms (MUSIC and ESPRIT) for both one and two dimensional cases and make relevant comparisons of these two algorithms. LMS algorithm is used for adaptive beam forming techniques for both of these DOA algorithms. The main goal of this paper is to design smart antennas for DECT system in WLL in 1.88 GHz—1.90GHz. One essential component of smart-antenna is its sensors or antenna elements. These antenna elements play an important role in shaping and scanning the radiation patterns and constraining the adaptive algorithm used by digital signal processor. Microstrip antennas gain popularity for its simple and inexpensive manufacturing using modern printed-circuit technology. This type of antenna is mechanically robust when mounted on rigid surfaces. As microstrip antenna has many advantages, we considered microstrip antenna for antenna array design. We designed a 64 element planar array antenna for the smart antenna system as planar array antenna could control the radiation pattern both azimuth and elevation directions. The performance of an adaptive antenna array is strongly affected by the electromagnetic characteristics of antenna array. An important electromagnetic characteristic of an antenna array is the mutual coupling between its elements [10]. There are several publications on the issue of mutual coupling effects for Adaptive arrays [10] − [12]. The presence of mutual coupling between the array elements degrade the array performance and reduce the speed of responses of an adaptive array [10]. So, we consider mutual coupling of antenna array elements to compensate the weights of Adaptive beamforming algorithm (Least Mean Square Algorithm).

II. Antenna Array Design

For microstrip antenna array we use corporate−feed network. The corporate–feed network is used to provide power splits of 2^n (i.e., n=2, 4, 8, 16, 32, etc.). This is accomplished by using either tapered lines or using quarter wavelength impedance transformers. With this method the designer has more control of the feed of each element both in amplitude and phase. Let us assume that we have $M \times N$ identical elements, M and N being even, with uniform spacing positioned symmetrical in the $xy-plane$. The array factor for this type of planar array with its maximum along θ_0 , ϕ_0, for an even number of elements in each direction can be written as [13]

$$[AF(\theta,\phi)]_{M\times N} = 4\sum_{m=1}^{M/2}\sum_{n=1}^{N/2} w_{mn}\cos[(2m-1)u]\cos[(2n-1)v] \quad (1)$$

$$u = \frac{\pi d_x}{\lambda}(\sin\theta\cos\phi - \sin\theta_0\cos\phi_0)$$

$$v = \frac{\pi d_y}{\lambda}(\sin\theta\sin\phi - \sin\theta_0\sin\phi_0)$$

where w_{mn} is the amplitude excitation of the individual elements. It is the w_{mn} that the adaptive beamforming algorithms adjust to place the maximum of the main beam toward the signal of interest and nulls toward the signal not of interest.

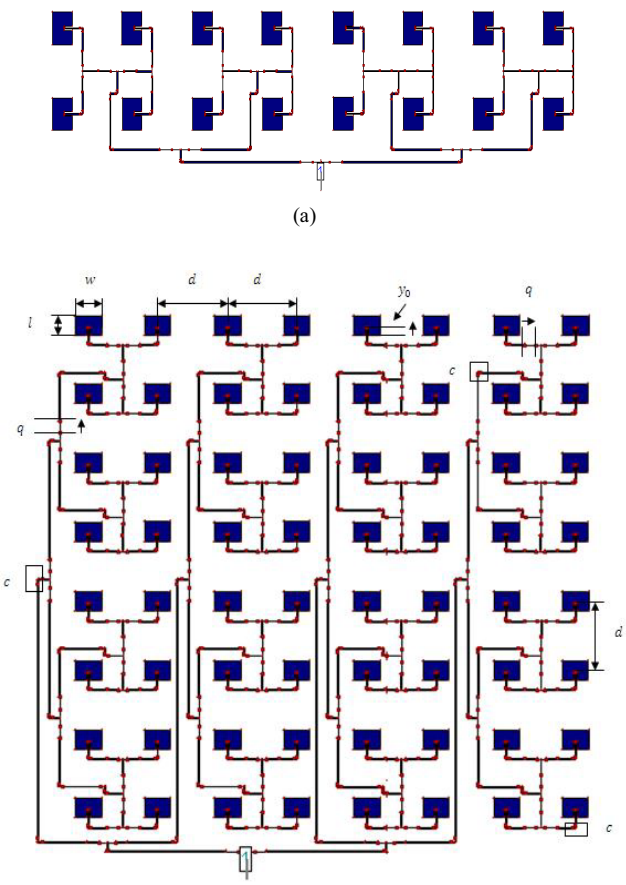
Substrate material of this antenna array is considered R03010 ($\epsilon_r = 10.2$), high dielectric constant is taken for size reduction because the operating frequency of the antenna is 1.88GHz − 1.90GHz. In these range of frequency antenna size is usually large unless high dielectric constant substrate is not used for antenna design. Recessed microstrip line feeding techniques is used as this gives a good impedance matching at inputs of the radiating elements. Feed networks in general have certain undesired characteristics that must be carefully monitored in order to minimize any adverse effects on array performance. These characteristics include conductor and dielectric losses, surface wave loss, and spurious radiation due to discontinuities such as bends, junctions and transitions. These losses constitute the overall insertion losses of the feed system affecting the maximum obtainable gain of the array [14]. For this simulated antenna, chamfered bend is used to compensate for excess capacitance [15]. Simulated antenna layout is shown in Fig.1 and dimensional parameters of the simulated antenna are given is Table 1. In Fig. 1(b) c indicates chamfered bend and antenna dimension is $700(l) \times 691(w) \times 1.58(h)mm$.

After Simulation we found that return loss is −33.10dB at 1.88GHz. Return loss of -10dB or below is found from 1.855GHz to 1.907 GHz, so it is suitable for use in 1.88/1.90GHz frequency band. 2.77% bandwidth is found for this antenna. Figure 2 shows the return loss curve. Side lobe level (SLL) of the antenna is more than 15dB lower than main lobe. Fig. 3 and 4 gives the radiation pattern of the simulated antenna. Fig. 5 gives the gain of the antenna at different frequencies. Simulated gain and directivity of the antenna is 21.0dBi and 23.39dBi respectively.

After the antenna array receives the incoming signals from all directions, the DOA algorithm determines the directions of all incoming signals based on the *time delay* which, for a $M \times N$ planar array can be computed using following equation [13]

$$\tau_{mn} = \frac{md_x\sin\theta\cos\phi + nd_y\sin\theta\cos\phi}{v_0} \quad (2)$$

where v_0 is the speed of light in free space.



(a)

(b)

Figure 1. Layout of the simulated antenna.

TABLE I.
DIMENSIONAL PARAMETERS OF THE ANTENNA.

w	l	d	q	y_0
34mm	25mm	88mm	15.6mm	10.13mm

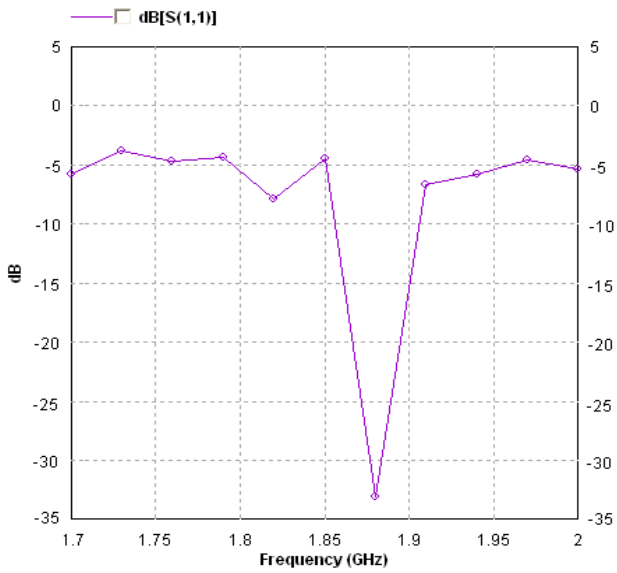


Figure 2. Return loss of the array antenna (64 elements).

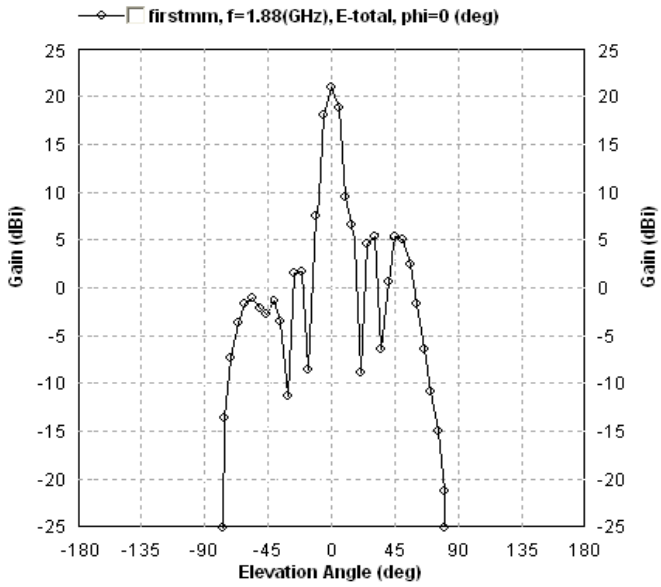


Figure 3. Radiation pattern of the array antenna.

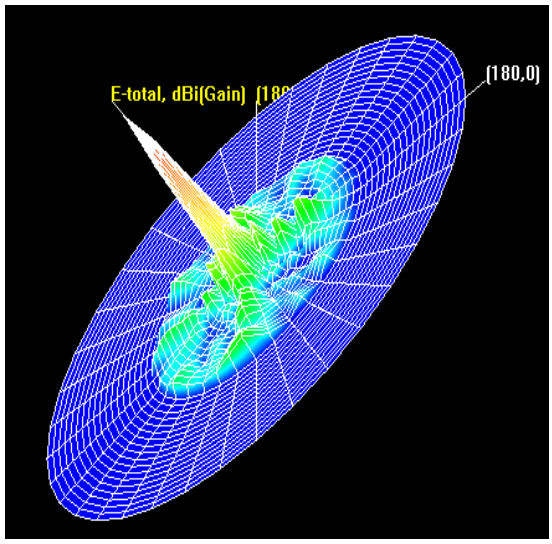


Figure 4. 3D Radiation pattern of the array antenna.

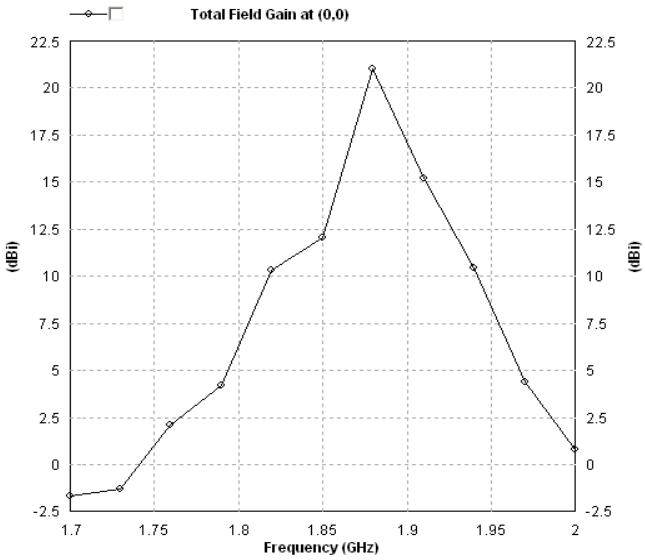


Figure 5. Simulated gain of the antenna at different frequencies.

III. DECT System in WLL And Rayleigh-Fadiang Channel

DECT system may have various different physical implementations depending on its actual use. Different DECT entities can be integrated into one physical unit; entities can be distributed, replicated etc. A good DECT system architecture model is given in ref. [3] and [16]. DECT system use GFSK (Gaussian Frequency Shift Keying) modulation techniques (BT=0.5).It's bit rate is 1.152Mbps, frame cycle time is 10ms and TDD duplex method is used. Wireless communication systems are characterized by time-varying multipath propagation channels, which are typically modeled as fading channels. In mobile radio channels, the Rayleigh distribution is commonly used to describe the statistical time varying nature of the received envelope of a flat fading signal, or the envelope of an individual multipath component [17]. Fig. 6 shows a typical Rayleigh fading envelope at 1890MHz, frequency within DECT range and receiver speed is considered $15\,km/h$. The BER (bit-error-rate) of GFSK modulation is evaluated over Rayleigh fading channel where received signal is corrupted by obstacles and causes multipath propagation as shown in Fig. 7. DECT has been used for Fixed Wireless Access as a substitute for copper pairs and provides a cost efficient means to establish the final drop in a public telecommunication network (Fig. 8). Wireless system always affected by multipath fading and interference; signal processing techniques can be used to improve this adverse condition. Smart antenna system is a solution to combat this problem. Smart antenna (adaptive array) work in the following way: First digital signal processor receives signals collected from each antenna element; it computes the direction−of−arrival (DOA) of the signal of interest (SOI). It then uses adaptive beamforming algorithms to produce a radiation pattern that focuses on the SOI, at the same time as tuning out any signal not of interest (SNOI). Fig. 9 shows the functional block diagram of a smart antenna system.

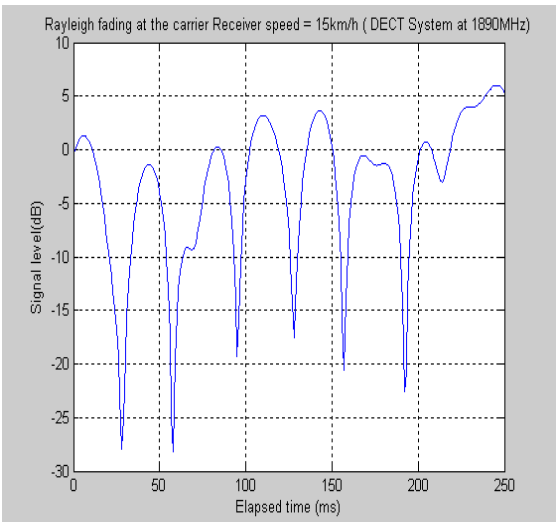


Figure 6. Rayleigh fading envelope at 1890MHz

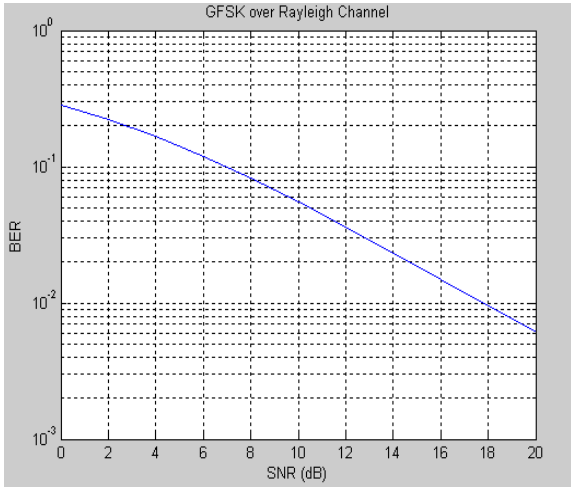


Figure 7. BER over Rayleigh fading channel for GFSK modulation

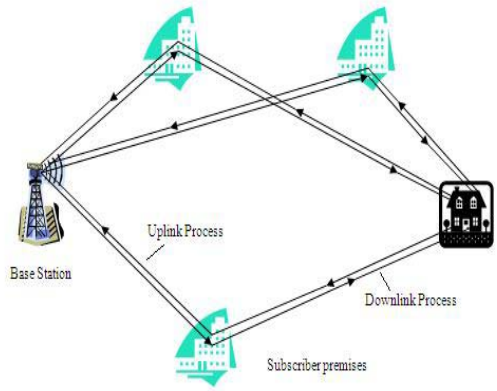


Figure 8. Block of DECT system for last mile solution.

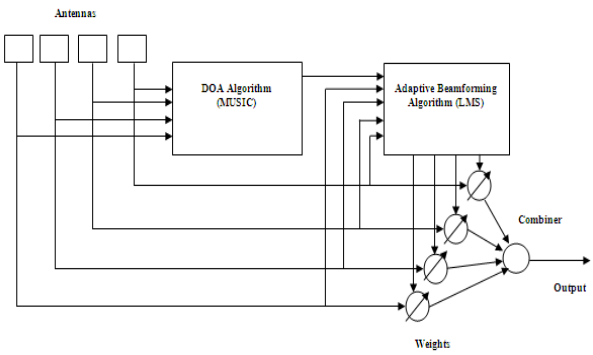


Figure 9. Block Diagram of a Smart Antenna System.

IV. DOA Estimation Algorithm

Subspace-based methods exploit the structure of the received data, resulting in an impressive improvement in resolution. **MU**ltiple **SI**gnal **C**lassification (MUSIC) algorithm fall into this category [13]. In our analysis we first consider MUSIC algorithm for DOA estimation using 8 element linear array (only azimuth angle). MUSIC is the first subspace-based DOA estimation approach (Schmit, 1986). MUSIC promises to provide unbiased estimates of the number of signals, the angle of arrival and the strengths of the waveforms. It makes the assumption that the noise in each channel is uncorrelated

making the noise correlation matrix diagonal and the incident signals may be somewhat correlated creating a nondiagonal signal correlation matrix [18]. MUSIC algorithm can be described in the following way:

Considering the number of signals is D and number of array elements is M. So the number of signal eigenvalues and eigenvectors is D and the number of noise eigenvalues and eigenvectors is $M-D$ $(M>D)$. Receive signal from M antennas can be written as follows:

$$x_k = [\ a(\theta_1)\ \ a(\theta_2)\ \ \ldots . a(\theta_D)\ ].\begin{bmatrix} s_1(k) \\ s_2(k) \\ s_3(k) \end{bmatrix} + \bar{n}(\mathrm{k})$$

$$= \ \bar{A}.\bar{s}(k) + \bar{n}(k) \tag{3}$$

$\bar{s}(k)$ is the vector of incident complex monochromatic signals at time k , $\bar{n}(k)$ is noise vector at each array element M, zero mean , variance σ_n^2 and $\bar{A}$ is $M \times D$ matrix of steering vectors $\bar{a}\theta_i$.

Antenna array output can be described as:

$$y_k = \bar{w}^T.\bar{x}_k \tag{4}$$

where $\bar{w} = [w_1\ \ w_2 \ldots\ldots . w_M\ \ ]^T$

Now $M \times M$ array correlation matrix can be written as follows,

$$\bar{\boldsymbol{R}}_{xx} = \boldsymbol{E}[\bar{x}.\bar{x}^H] = \boldsymbol{E}[(\bar{A}\bar{s} + \bar{n})(\bar{s}^H\bar{A}^H + \bar{n}^H)]$$
$$= \bar{A}\boldsymbol{E}[\bar{s}.\bar{s}^H]\bar{A}^H + \boldsymbol{E}[\bar{n}.\bar{n}^H] = \bar{A}\bar{\boldsymbol{R}}_{\boldsymbol{ss}}\bar{A}^H + \bar{\boldsymbol{R}}_{nn}$$
$$= \bar{A}\bar{\boldsymbol{R}}_{\boldsymbol{ss}}\bar{A}^H + \sigma_n^2\bar{\boldsymbol{I}} \tag{5}$$

where $\bar{\boldsymbol{R}}_{\boldsymbol{ss}} = D \times D$ source correlation matrix, $\bar{\boldsymbol{R}}_{\boldsymbol{nn}} = \sigma_n^2\bar{\boldsymbol{I}} = M \times M$ noise correlation matrix and $\bar{\boldsymbol{I}} = N \times N$ identity matrix. After this we can write $M \times (M-D)$ dimensional subspace spanned by the noise eigenvectors that is as follows,

$$\bar{\boldsymbol{E}}_{\boldsymbol{N}} = [\bar{e}_1\ \ \ \bar{e}_2\ \ \ldots\ldots\ \bar{e}_{M-D}] \tag{6}$$

The noise subspace eigenvectors are orthogonal to array steering vector at angles of arrival θ_1, θ_2 ,...., θ_D. At last MUSIC pseudo spectrum is defined as,

$$\boldsymbol{P}_{\boldsymbol{MU}}(\theta) = \frac{1}{|\bar{a}(\theta)^H\bar{\boldsymbol{E}}_{\boldsymbol{N}}\bar{\boldsymbol{E}}_{\boldsymbol{N}}^{\boldsymbol{H}}\bar{a}(\theta)|} \tag{7}$$

The MUSIC algorithm in general can apply to any arbitrary array regardless of the position of the array elements [18]. Now we consider signal received from both azimuth angle and elevation angle. In this case signals are collected by an array made up of $M \times N$ antennas with direction of arrival (θ_D, ϕ_D). θ_D is the azimuth angle and ϕ_D is the elevation angle.

Receive signal from $M \times N$ antennas can be written as follows:

$$x_k = [\ a(\theta_1, \phi_1)\ \ a(\theta_2, \phi_2)\ \ \ldots . a(\theta_D, \phi_D)\ ]\ .\begin{bmatrix} s_1(k) \\ s_2(k) \\ s_3(k) \end{bmatrix} + \bar{n}(\mathrm{k})$$
$$= \bar{A}.\bar{s}(k) + \bar{n}(k) \tag{8}$$

The antenna array output can be described in the following form:

$$y_k = \bar{w}^T.\bar{x}_k \tag{9}$$

where $\bar{w} = [w_1\ \ w_2 \ldots\ldots . w_M\ \ ]^T$

The angle of arrival can be describe using following equations [19],

$$a(\theta_D, \phi_D) = [1\ e^{-\mathrm{jk}(\cos\theta_\mathrm{k}\ .\ \cos\phi_\mathrm{k}\ +\ \cos\theta_\mathrm{k}.\ \sin\phi_\mathrm{k})\ldots\ldots}$$
$$e^{-\mathrm{jk}[(M-1)\cos\theta_k\ .\ \cos\phi_k + (N+1)\cos\theta_k.\sin\phi_k]}\ ] \tag{10}$$

where, M and N are the element number for x and y axis respectively. Applying the same estimation method that is used for one dimension, the pseudo spectrum of MUSIC algorithm is as follows:

$$\boldsymbol{P}_{\boldsymbol{MU}}(\theta, \phi) = \frac{1}{|\bar{a}(\theta,\phi)^H\bar{\boldsymbol{E}}_{\boldsymbol{N}}\bar{\boldsymbol{E}}_{\boldsymbol{N}}^{\boldsymbol{H}}\bar{a}(\theta,\phi)|} \tag{11}$$

We used MUSIC algorithm for uplink condition. We assumed that a smart antenna is only employed at the base station, and not at the subscriber unit. Consider signal arrival from four different directions in different angles; these angles are randomly taken as the incoming signals are random in nature. Fig .10 illustrates the pseudo spectrum of MUSIC with four different angle−of −arrival for 8 element linear array antenna with SNR value of 18dB. Fig. 11 shows the possible radiation pattern of the beamformer of that antenna considering azimuth angle. Fig. 12 gives the MUSIC pseudo spectrum for 8×8 planar array antenna considering both azimuth and elevation angles with SNR value of 18dB. Fig. 13 illustrates the possible radiation pattern of that antenna. We see that in both cases some phase errors are introduced in the radiation pattern of the beamformer (Fig. 11 and Fig. 13). Mutual coupling effect between the antenna element and improper calculation of weights in LMS algorithm can cause this problem. From Fig 10. We can realize that if two sources (angle-of-arrival at 15.4105^0 and 13.4023^0) are very close then MUSIC algorithm is unable to resolve this sources.

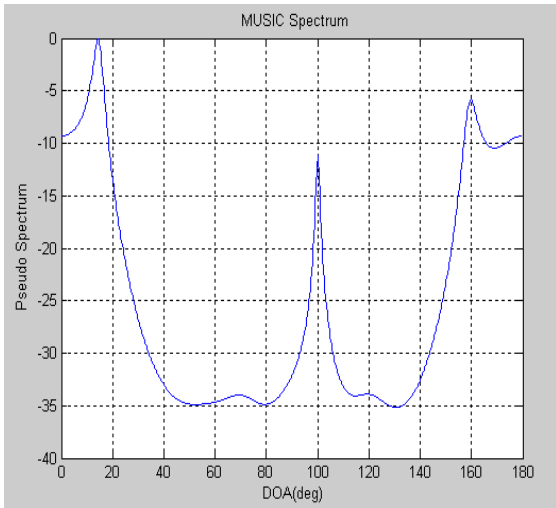


Figure 10. Pseudo spectrum of MUSIC with angle of arrival (azimuth direction) at 15.4105^0, 13.4023^0, 100.1832^0 , 160.3601^0 and SNR is 18dB.

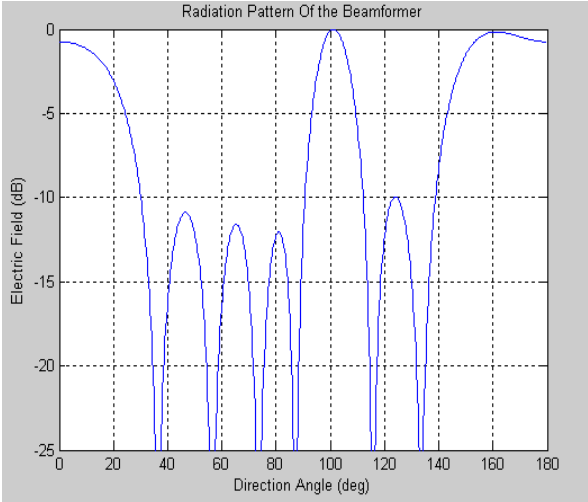
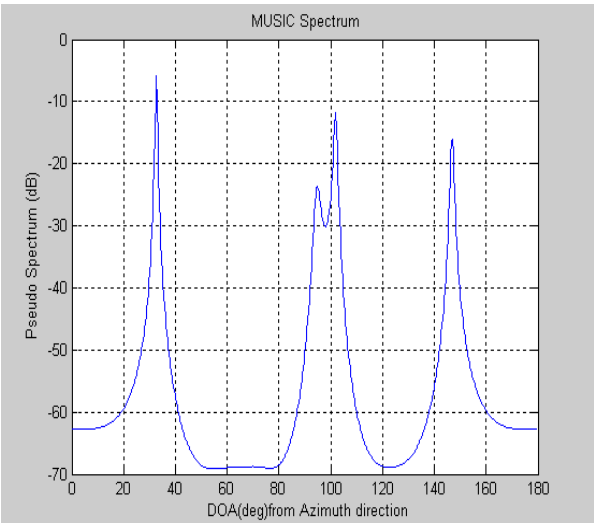
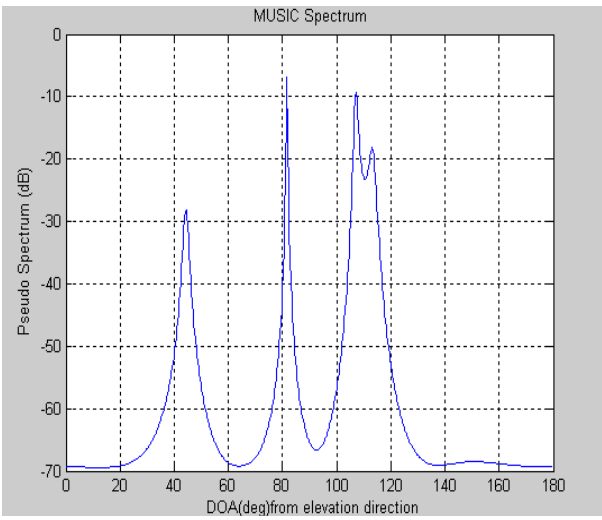


Figure 11. Possible Radiation pattern of the beamformer for signal arrival (azimuth direction) at 15.4105^0, 13.4023^0, 100.1832^0, 160.3601^0 and SNR is 18dB.

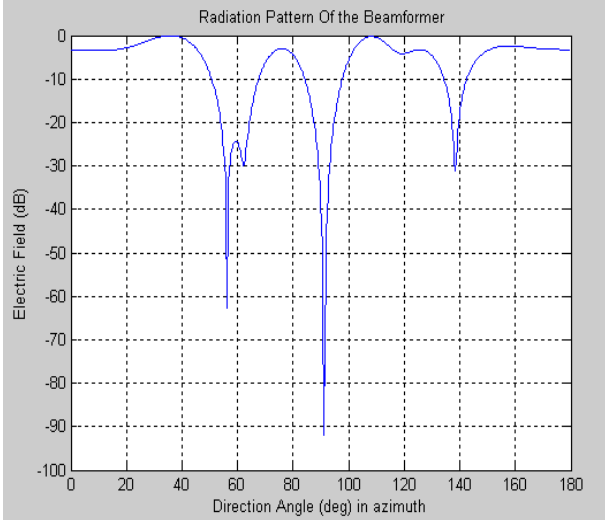


(a)

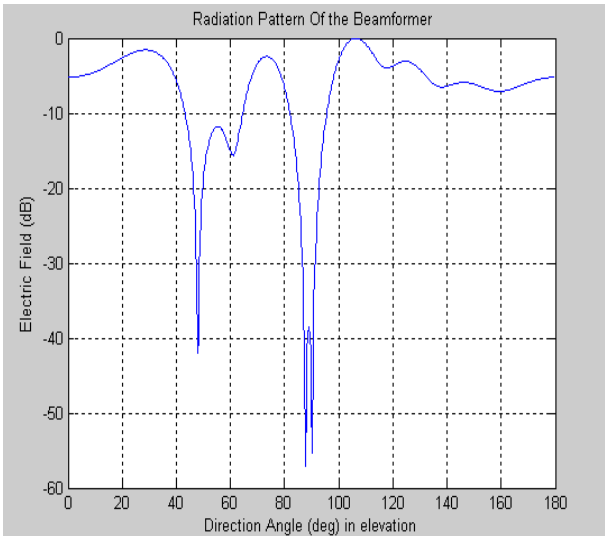


(b)

Figure 12. Pseudo spectrum of MUSIC with angle of arrival at $(95.6564^0, 81.6510^0)$, $(147.6541^0, 113.4150^0)$, $(36.1061^0, 44.1702^0)$, $(101.4183^0, 107.0332^0)$ and SNR is 18dB.



(a)



(b)

Figure 13. Possible Radiation pattern of the beamformer for signal arrival at $(95.6564^0, 81.6510^0)$, $(147.6541^0, 113.4150^0)$, $(36.1061^0, 44.1702^0)$, $(101.4183^0, 107.0332^0)$ and SNR is 18dB.

Another subspace based method is ESPRIT (*Estimation of Signal Parameters via Rotational Invariance Techniques*) which gives some advantages over MUSIC algorithm like less computationally intensive, requires much less storage, does not involve an exhaustive search through all possible steering vectors to estimate the direction-of-arrival. First we consider the ESPRIT algorithm for one dimensional condition (azimuth angle). The goal of the ESPRIT techniques is to exploit the rotational invariance in signal subspace which is created by two arrays with a translational invariance structure [18]. As like MUSIC, ESPRIT assumes that the number of signal D is less that number of antenna element M $(M > D)$. The idea behind ESPRIT is to divide the array in two equivalent subarrays separated by a known displacement d. Here we consider eight element linear arrays that are composed of two identical five element subarrays or two doublets. The signals induced

on each of the arrays can be described by following equations:

$$\bar{x}_1(k) = [\ a(\theta_1)\ \ a(\theta_2)\ \ ....a(\theta_D)\ ].\begin{bmatrix} s_1(k) \\ s_2(k) \\ \vdots \\ . \\ s_3(k) \end{bmatrix} + \bar{n}_1(k)$$

$$= \ \bar{A}_1.\bar{s}(k) + \bar{n}_1(k) \qquad (12)$$

Similarly,

$$\bar{x}_2(k) \ = \ \bar{A}_2.\bar{s}(k) + \bar{n}_2(k)$$

$$= \ \bar{A}_1.\bar{\Phi}.\bar{s}(k) + \bar{n}_2(k) \qquad (13)$$

where $\bar{\Phi} = diag\{e^{jkd\sin\theta_1}, e^{jkd\sin\theta_2}, \dots, e^{jkd\sin\theta_D}\}$ = a $D \times D$ diagonal unitary matrix with phase shifts between the doublets for each AOA.

$\bar{A}_i$ = Vandermonde matrix of steering vectors for subarrays $i = 1,2$

The complete receive signal can be expressed by following equation:

$$\bar{x}(k) = \begin{bmatrix} \bar{x}_1(k) \\ \bar{x}_2(k) \end{bmatrix} = \begin{bmatrix} \bar{A}_1 \\ \bar{A}_1.\bar{\Phi} \end{bmatrix}.\ \bar{s}(k)\ + \begin{bmatrix} \bar{n}_1(k) \\ \bar{n}_2(k) \end{bmatrix} \qquad (14)$$

Now the correlation matrix for the complete array is given by

$$\bar{\boldsymbol{R}}_{xx} = \boldsymbol{E}[\bar{x}.\bar{x}^H] = \bar{A}\bar{\boldsymbol{R}}_{\boldsymbol{ss}}\bar{A}^H + \sigma_n^2\bar{\boldsymbol{I}} \qquad (15)$$

whereas the correlation matrices for the two subarrays are given by

$$\bar{\boldsymbol{R}}_{11} = \boldsymbol{E}[\bar{x}_1\bar{x}_1^H] = \bar{A}\bar{\boldsymbol{R}}_{\boldsymbol{ss}}\bar{A}^H + \sigma_n^2\bar{\boldsymbol{I}} \qquad (16)$$

and

$$\bar{\boldsymbol{R}}_{22} = \boldsymbol{E}[\bar{x}_2\bar{x}_2^H] = \bar{A}\bar{\boldsymbol{R}}_{\boldsymbol{ss}}\bar{A}^H + \sigma_n^2\bar{\boldsymbol{I}} \qquad (17)$$

using the above two equations we can construct the signal subspaces $\bar{E}_1$ and $\bar{E}_2$. The entire array signal subspace is $\bar{E}_x$. $\bar{E}_x$ is an $M \times D$ matrix composed of signal eigenvectors. $\bar{E}_1$ can be constructed by selecting the first $M/2 + 1$ rows of $\bar{E}_x$. $\bar{E}_2$ can be constructed by selecting the last $M/2 + 1$ rows of $\bar{E}_x$. Now a $2D \times 2D$ matrix can be formed using the signal subspace that is as follows:

$$\bar{C} = \begin{bmatrix} \bar{E}_1^H \\ \bar{E}_2^H \end{bmatrix} [\bar{E}_1\ \bar{E}_2] = \ \bar{E}_c\bar{\Lambda}\ \bar{E}_c^H \qquad (18)$$

where the matrix $\bar{E}_c$ is from the eigenvalue decomposition (EVD) of $\bar{C}$ such that $\lambda_1 \geq \lambda_2 \geq \cdots \geq \lambda_{2D}$ and $\bar{\Lambda} = diag\{\ \lambda_1, \lambda_2, \dots \lambda_{2D}\}$

Performing the eigendecomposition we can construct the matrix $\bar{E}_c$ in following way,

$$\bar{E}_c = \begin{bmatrix} \bar{E}_{11} & \bar{E}_{12} \\ \bar{E}_{21} & \bar{E}_{22} \end{bmatrix} \qquad (19)$$

The rotational operator $\bar{\psi}$ can be expressed by following equation,

$$\bar{\psi} = -\bar{E}_{12}\bar{E}_{22}^{-1} \qquad (20)$$

After calculation of the eigenvalues of $\bar{\psi}$, $\lambda_1, \lambda_{2,\dots,}\lambda_D$ we can estimate the angle of arrival,

$$\theta_i = \sin^{-1}\left(\frac{\arg(\lambda_i)}{kd}\right) \qquad (21)$$

where $\lambda_i = |\lambda_i|e^{jarg(\lambda_i)}$ and $i = 1,2,\dots,D$

We now consider *ESPRIT* algorithm for two dimensional conditions (both azimuth angle and elevation angle). In reference [20], the unitary ESPRIT algorithm was presented for DOA estimation for uniform rectangular arrays. For a uniform rectangular array of $N \times M$ elements lying in the x-y plane and equispaced by Δ_x in the x direction and Δ_y in the y direction. The DOA of the source is specified by the pair (u, v) , where $u = \sin\theta\ \cos\phi$ and $v = \sin\theta\ \sin\phi$ are the direction cosines with respect to the x and y axes, respectively. When a narrow band source impinges on the array from the direction(u, v), the phase shifts between successive elements along the x and y axes are $\mu = 2\pi\frac{\Delta_x}{\lambda}u$ and $v = 2\pi\frac{\Delta_y}{\lambda}v$ respectively. μ and v lie in the range $[-\pi, \pi]$ when $\Delta_x = \Delta_y = \lambda/2$. The array output can be modeled as $x(t) = As(t) + n(t)$, where $x(t)$ is the NM vector formed by stacking the columns of uniform rectangular array outputs, A is the $NM \times d$ DOA matrix (assuming d incident sources) , $s(t)$ is the vector of signal complex envelopes at the origin and $n(t)$is the stacked noise vector [21]. Here A can be expressed as [20],

$$A(\mu, v) = \ a_N(\mu)a_M^T(v) \qquad (22)$$

$$a_M(v) = \left[e^{-j\left(\frac{M-1}{2}\right)v}, \dots e^{-jv}, 1, e^{jv}, \dots, e^{j\left(\frac{M-1}{2}\right)v}\right]^T$$

Premultiplying $A(\mu, v)$ by Q_N^H and post-multiplying by Q_M^* , creates the $N \times M$ real-valued manifold [20],

$$D(\mu, v) = Q_N^H A(\mu, v)Q_M^* \ = Q_N^H\ \ a_N(\mu)a_M^T(v)Q_M^* \ = d_N(\mu)d_M^T(v) \qquad (23)$$

$$d_M(v) = Q_M^H a_M(v) = \sqrt{2} \times \left[\cos\left(\frac{M-1}{2}v\right), \dots \cos(v), 1/\sqrt{2}, -\ \sin\left(\frac{M-1}{2}v\right), \dots, -\sin(v)\right]$$

where Q_M^H is a sparse unitary matrix that transforms $a_M(v)$ into an $M \times 1$ real−valued manifold.

$D(\mu, v)$ satisfies the following equation,

$$\tan\left(\frac{\mu}{2}\right)K_1D(\mu, v) = K_2D(\mu, v) \qquad (24)$$

where $K_1 = \mathcal{R}e\{Q_{N-1}^H J_2\ Q_N\}$ and $K_2 = Im\{Q_{N-1}^H J_2\ Q_N\}$. Now, we find that the $NM \times 1$ real-valued manifold in vector from, $d(\mu, v) = vec[D(\mu, v)]$ satisfies the following equation ($vec[D(\mu, v)]$ represents vectorization of D)

$$\tan\left(\frac{\mu}{2}\right)K_{\mu 1}d(\mu, v) = K_{\mu 2}d(\mu, v) \qquad (25)$$

Where $K_{\mu 1}$ and $K_{\mu 2}$ are the $(N-1)M \times NM$ matrices

$$K_{\mu 1} = \ I_M \otimes K_1 \ \text{ and } K_{\mu 2} = \ I_M \otimes K_2 \qquad (26)$$

Here $\otimes$ denotes the Kronecker matrix product. Now, we can write the following equation,

$$\tan\left(\frac{v}{2}\right) D(\mu,v)K_3^T = D(\mu,v)K_4^T \quad (27)$$

where $K_3 = \mathcal{R}e\{Q_{M-1}^H J_2\, Q_M\}$ and $K_4 = Im\{Q_{M-1}^H J_2\, Q_M\}$ using the vec operator, we find the $d(\mu,v)$ satisfies

$$\tan\left(\frac{v}{2}\right) K_{v1} d(\mu,v) = K_{v2} d(\mu,v) \quad (28)$$

where K_{v1} and K_{v2} are the $N(M-1)\times NM$ matrices

$$K_{v1} = K_3 \otimes I_N \quad \text{and} \quad K_{v2} = K_4 \otimes I_N \quad (29)$$

Consider the $NM \times d$ real-valued DOA matrix $D = [d(\mu_1,v_1),\ldots,d(\mu_d,v_d)]$, where $d(\mu,v) = vec[D(\mu,v)]$ D satisfies the following equation,

$$K_{\mu1} D\Omega_\mu = K_{\mu2} D \quad (30)$$

where, $\Omega_\mu = diag\{\tan\left(\frac{\mu_1}{2}\right),\ldots,\tan(\frac{\mu_d}{2})\}$; similarly D satisfies the following equation,

$$K_{v1} D\Omega_v = K_{v2} D \quad (31)$$

where, $\Omega_v = diag\{\tan\left(\frac{v_1}{2}\right),\ldots,\tan(\frac{\mu_d}{2})\}$.

Now , $E_s = DT$ where T is an unknown $d \times d$ real-valued matrix. Substituting $D = E_s T^{-1}$ into the (30) and (31) yields the signal eigenvector relations

$$K_{\mu1} E_s \Psi_\mu = K_{\mu2} E_s \;\; ; \; \Psi_\mu = T^{-1}\Omega_\mu T \quad (32)$$

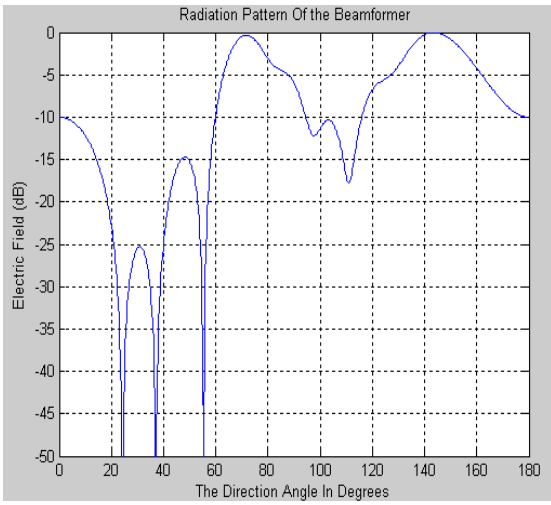
$$K_{v1} E_s \Psi_v = K_{v2} E_s \;\; ; \; \Psi_v = T^{-1}\Omega_v T \quad (33)$$

Thus, we can write $\Psi_\mu + j\Psi_v$ in following form,

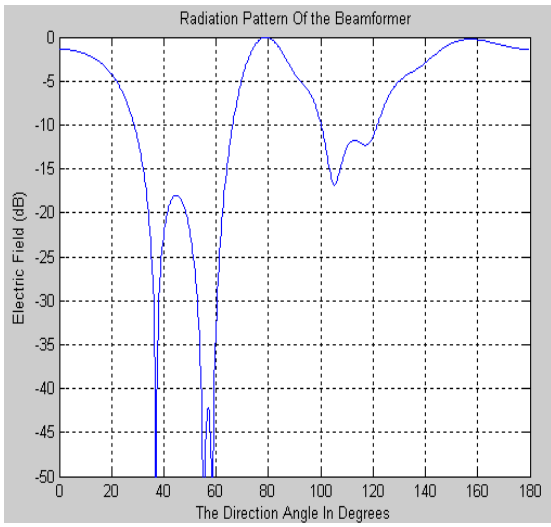
$$\Psi_\mu + j\Psi_v = T^{-1}\{\Psi_\mu + j\Psi_v\}T \quad (34)$$

The spatial frequency estimates are,
$\mu_i = 2\tan^{-1}(\mathcal{R}e\{\lambda_i\})$ and $v_i = 2\tan^{-1}(Im\{\lambda_i\})$; $i = 1,\ldots d$ Note that $2-$D unitary ESPRIT provides closed-form, automatically paired $2-$D angle estimates as long as the spatial frequency coordinate pairs (μ_i, v_i) $i = 1,\ldots,d$ are unique [20].

We simulated ESPRIT algorithm for both one and two dimensional cases. Fig. 14 gives the possible radiation pattern of the beamformer (azimuth angle) for 8 element linear array antenna. Four random signals with different angle−of−arrivals are considered. Fig. 15 illustrates the possible radiation pattern of the beamformer using $2-$D ESPRIT algorithm considering both azimuth and elevation angles also in this case we consider four random signals with different directions. We simulated the program many times by taking random incoming signals and Fig. 15 is one of the simulation results. Simulation depicts that the phase errors in radiation pattern of the beamformer is small for both $1-$D and $2-$D ESPRIT algorithm than MUSIC algorithm although same adaptive beamforming algorithm (LMS algorithm) have been used. The most important thing that we investigated was the simulation time for $2-D$ ESPRIT and $2-D$ MUSIC algorithm; MUSIC algorithm requires much time for simulation than ESPRIT for the same number of random incoming signals (angle−of−arrivals). So ESPRIT algorithm can save valuable signal processing time for Smart antenna system.

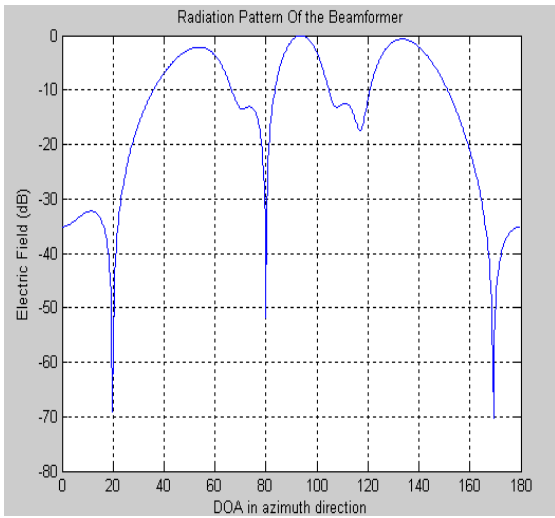


(a)

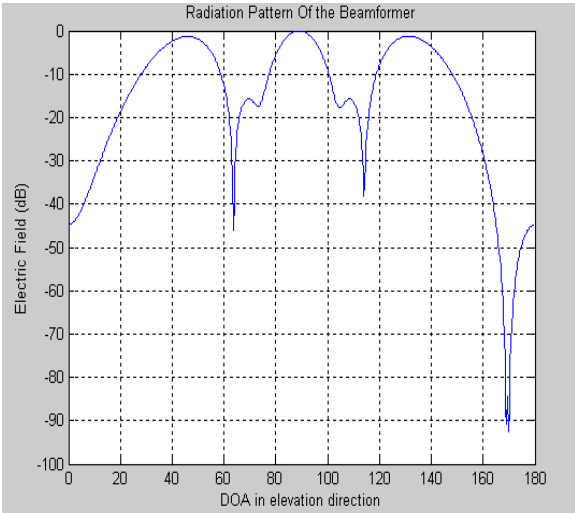


(b)

Figure 14. Possible radiation pattern of the beamformer for signal arrival at $74.2600^0, 140.4647^0, 158.4201^0, 10.9470^0$ and SNR is 18dB.



(a)



(b)

Figure 15. Possible radiation pattern of the beamformer for signal arrival at $(53.1242^0, 51.4569^0)$, $(92.2425^0, 41.40^0)$, $(125.5253^0, 90.0^0)$, $(136.5021^0, 130.1212^0)$ and SNR is 18dB.

V. Adaptive Beamforming Techniques

In adaptive beamforming, the target is to adapt the beam by adjusting the amplitudes and phases of signals such that an enviable pattern is formed. One of the simplest algorithms that are commonly used to adapt the weights is the Least Mean Square (LMS) algorithm. The LMS algorithm is a low complexity algorithm that requires no direct matrix inversion and no memory. It is an approximation of the steepest descent method using an estimator of the gradient instead of the actual value of the gradient, since computation of the actual value of the gradient is impossible because it would require knowledge of the incoming signals. As a result, at each iteration in the adaptive process, the estimate of the gradient is as follows [13].

$$\hat{\nabla}[J(\boldsymbol{w})]_k = \begin{bmatrix} \frac{\partial J(\boldsymbol{w})}{\partial w_0} \\ \vdots \\ \frac{\partial J(\boldsymbol{w})}{\partial w_L} \end{bmatrix} \tag{35}$$

Where the $J(\boldsymbol{w})$ is the cost function. We can obtain this by following mathematical analysis,
x_k is a vector from each antenna array element at time step k [9],

$$x_k = [x(k)_1\ x(k)_2 \ldots\ldots\ x(k)_L]^T \tag{36}$$

respectively, we also get a weight vector w

$$w = [\, w_1\ w_2 \ldots\ldots\ w_L\,] \tag{37}$$

Then the output $y(k)$ for time step k is

$$y(k) = w^H\ x(k) \tag{38}$$

ε_k is the error between the desired signal d_k and the output signal of the array $y(k)$, can be expressed as,

$$\varepsilon_k = d_k - \boldsymbol{w}^H x_k \tag{39}$$

MSE (Mean Square Error) based cost function,

$$J_{MSE}(E[\varepsilon_k^2]) = d_k^2 - 2\boldsymbol{w}^H E[d_k \boldsymbol{x}_k] + \boldsymbol{w}^H E[\boldsymbol{x}_k \boldsymbol{x}_k^H]\boldsymbol{w} \tag{40}$$

$$J_{MSE}(E[\varepsilon_k^2]) = d_k^2 - 2\boldsymbol{w}^H r_{xd} + \boldsymbol{w}^H R_{xx}\boldsymbol{w} \tag{41}$$

where $\boldsymbol{r}_{xd} = E[d_k \boldsymbol{x}_k]$ and $\boldsymbol{R}_{xx} = E[\boldsymbol{x}_k \boldsymbol{x}_k^H]$ .Using steepest descent method, the iterative equation updates weights at each iteration; this can be expressed by following equation,

$$\boldsymbol{w}_{k+1} = \boldsymbol{w}_k - \mu\bar{\nabla}[J(\boldsymbol{w})]_k \tag{42}$$

where μ is the step size related to the rate of convergence. This simplifies the calculation to be performed considerably and allows LMS algorithm to be used in real-time applications. LMS algorithm minimizes the MSE (Mean Square Error) cost function and it solves the Wiener-Hopf equation iteratively without the need for matrix inversion. The Wiener-Hopf equation is as follows,

$$\boldsymbol{w}_{opt} = \boldsymbol{R}_{xx}^{-1}\boldsymbol{r}_{xd} \tag{43}$$

The LMS algorithm computes the weights iteratively as,

$$\boldsymbol{w}_{k+1} = \boldsymbol{w}_k + 2\mu x_k(d_k - x_k^T \boldsymbol{w}_k) \tag{44}$$

In order to assure convergence of the weights,$\boldsymbol{w}_k$, the step size μ is bounded by the following condition

$$0 < \mu < \frac{1}{\lambda_{max}} \tag{45}$$

where λ_{max} is the maximum eigenvalue of the covariance matrix, R_{xx} given is equation (5). The main disadvantage of the LMS algorithm is that it tends to converge slowly, particularly in noisy environment. Fig. 16 shows the simulated result of actual and estimated system output using LMS algorithm. Sixty-four weights are considered for simulation. Fig. 17 illustrates the error curve that gives error between the desired signal and output signal of the array. At last Fig. 18 describes the comparison of actual weights and estimated weights. The complex weights $\boldsymbol{w}_k$ in equation (44) are the ideal weights if mutual coupling is not considered. These weights should be compensated according to mutual coupling in order to get better far field patterns. Compensation for mutual coupling can be accomplished by simply multiplying the received signal x_k as given in equation (36) by the inverse mutual coupling matrix C^{-1} [11]. The mutual coupling matrix can be expressed using following equations [11]

$$C = I + \boldsymbol{Z}Z_L^{-1} \tag{46}$$

where I is the identity matrix, $\boldsymbol{Z}$ is the impedance matrix (analysis was given in ref. [10]) and Z_L is the load impedance (i.e. 50Ω). Data of the impedance matrix are obtained form antenna array simulations using *Zeland IE3D* and these results are used in mutual coupling matrix (equation (46)) calculation. We can express the received and compensated signals using the following equation,

$$x_k^d = C^{-1}x_k \tag{47}$$

x_k^d denote the received and compensated signals.

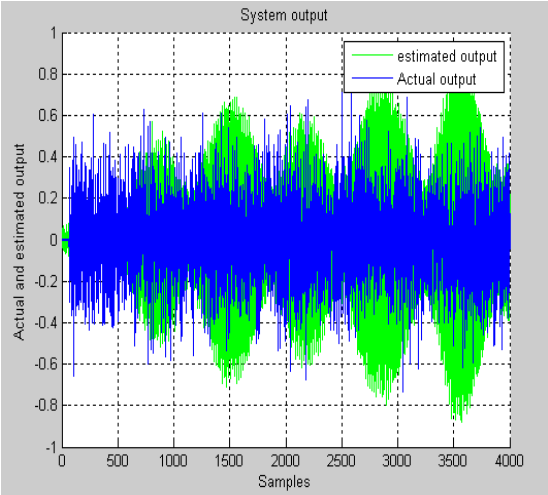


Figure 16. System output comparison of Actual value and estimated value.

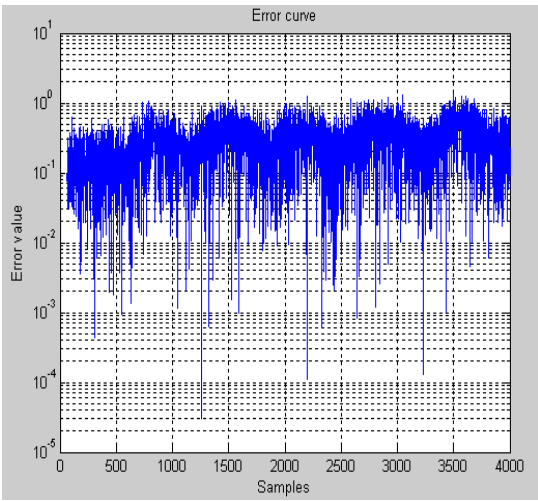


Figure 17. Error value between actual and estimated output.

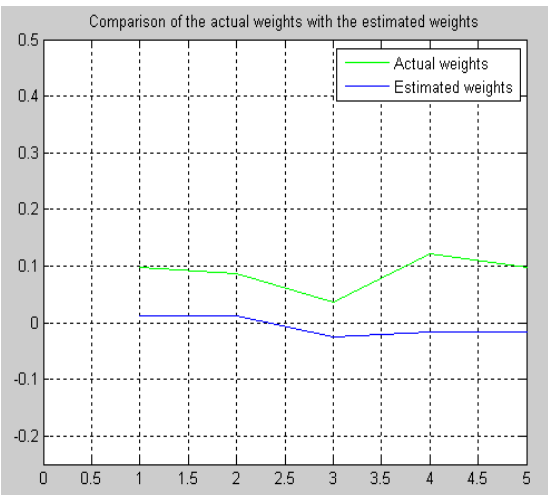


Figure 18. Comparison of Actual weights and the estimated weights, iteratively computes using equation (44).

VI. Conclusion

A smart antenna system has been designed to minimize the adverse effect of time-varying multipath propagation channel and improvement of the signal communication link for DECT RBS in WLL system. For signal processing we exploit MUSIC and ESPRIT algorithm considering both 1-D (azimuth angle) and 2-D (azimuth and elevation angle) angle−of −arrivals. LMS algorithm is used for adaptive beamforming as it requires low complexity. Using LMS algorithm we calculate the weights for adaptive beamforming; in last portion we show the differences between actual weights and the estimated weights regarding LMS algorithm. Mutual coupling is considered as this gives more realistic results. At last we can conclude that for same input signal (angle −of−arrival) and antenna array ESPRIT gives better performance than MUSIC algorithm. In future we want to work on the improvement of bit error rate (BER) using smart antenna for WLL and improvement of network throughput. Simulation of a 64 element microstrip antennas is very time consuming specially using low processing speed computer, Intel core 2 due CPU T8100 @ 2.10GHz , 2.09GHz is used for this task. Only 11 frequency points are considered from 1.7-2.0GHz for processing as more frequency points require more time. In simulation each frequency point takes approximately 2800sec and a total of 9 hours for 11 frequency points.

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