Smart Antenna for Wireless Communication Systems using Spatial Signal Processing

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Abstract —To realize smart antennas necessity with wideband in wireless communication systems, three main fundamental approaches are, (i) space-time signal processing, (ii) spatial-frequency signal processing (filtering of signals that overlay with noise in space and frequency), and (iii) spatial signal processing (beamforming). This research work deals with the spatial signal processing (beamforming) due to its spatial access to the radio channel through diverse approaches. This is based on directional specifications of the second order spatial analysis of communication radio channel. Space-time processing decreases the interference and improves the desired signal. In addition to beamforming, this research work also estimate the DOA arrays of smart antenna based on uniform linear array. Delay-and-sum beamformer response tuned to $\vec{\alpha}$ for monochromatic plane has been established in this work. The results for the analysis and simulations shown validate the benefits of the proposed technique.

Index Terms—Antenna arrays, beamforming, DOA estimation, smart antenna, spatial filtering, spatial signal processing, MUSIC

I. Introduction

Recently, smart antenna systems have drawn the attention of academia and industry because of the attached spatial filtering capacity of the digital signal processors in wireless communications system [1]. They can concurrently increase the functional received signal from a particular direction and suppressing undesired interference signals and noise in another direction [2], [3]. Hence, this improves power efficiency and capacity of the system. Thus, reducing the overall cost of the system equipment. Popular research on smart antennas considered narrow band signals. For a wide bandwidth smart antenna systems to be used [4], [5], the radiation beampattern of adaptation should consider the frequency dependency phase shift of the inter-element spacing of the antennas. The three major methods to implement smart antennas with wideband-width are:

(i) Signal processing in space and time (spatiotemporal beamformer): this system comprises of array of antennas/sensors and tapped-delay line at respective division of the antenna arrays to sort out the signal received in time domain [6].

(ii) Space-frequency signal processing technique: this is another method of accomplishing wide bandwidth beam-formation devoid of tapped-delay lines. The received signal can be broken down into non-overlapping narrowband element through the aid of band-pass filter, and (iii) Conventional (wide-bandwidth spatial beamformer or fully spatial signal processing) method: This technique uses 2-D Inverse Discrete Fourier Transform (IDFT) to calculate the weighting coefficients of signal processing [6]. Signal processing in spatial form or beamforming is the utmost appealing one, and the processed signals are only in space domain [7], [8].

Smart antennas are sophisticated controlling systems that process signals stimulated on sensors (antennas) array [9]-[17] in adaptive manner. By dynamically adapting the beam pattern function of the sensors array using the signal processing [11], smart antennas can expand the network capacity and improve the interference rejection [12]-[14]. The sensors of antennas array structures are separated spatially. The major characteristics of smart antennas consist of:

(i) Capability to evaluate the direction of an arriving signals. This is normally referred to as direction of arrival (DOA), and

(ii) Capability to regulate the beam radiation pattern of antenna arrays [15].

Due to the spatial separation capacity of smart antenna, it offers the spatial division of various mobile users to a co-channel by means of Spatial Division Multiple Access (SDMA) technique, thus increasing the system capability [14]. Smart antenna uses complex signal-processing algorithms to:

(i) Continually separate among signal of interest, interfering signals and multipath, and

(ii) Compute their directions of arrival (DOA) [9].

Various researchers regarded smart antennas as one of the essential components in meeting the service needs of spectral efficiency and spatial processing in modern generation mobile network [15]-[17]. Using a spatial signal processing, the signals received are first converted to base band and then sampled. One of the advantages of spatial signal processing is that wideband beamforming can be successfully performed without tapped-delay lines or frequency filters. This can also be referred to as wideband spatial beamformer. Spatial signal processing is an adaptable method for improvement of Signal of Interest (SOI) while reducing interference signals and noise in the array of sensor's output [2].

Uthanksul and Bialkowski [4], a smart antenna through capacity of beam steering in azimuth angle over a

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wideband frequency by means of only spatial signal processing was validated. The wideband smart antennas were designed through filters and tapped-delay networks. The variable beam pattern radiation in the azimuth over a wideband of frequency was provided through a two-dimensional (2-D) antenna array entailing a wideband antenna elements aligned horizontally in a rectangular framework. The arrangement of the array weighing relied on the Inverse Discrete Fourier Transform (IDFT) technique. Wide bandwidth signals was not fully achieved through this method. Omnidirectional strategies were used in ref. [15], which has a direct and adverse impact on spectral efficiency, this limits frequency reuse. Similarly, mobile network systems adopting smart antenna for spatial-processing was presented in ref. [15], and an increase in capacity was accomplished.

A synopsis of array signal processing techniques for narrow-band signals was established in ref. [18]. In general, processing of array signal is useful in detection and elimination of difficulties being encountered when a desired signal is taken, interference and noise may likely occurs. *Sarkar and Adve* [3] uses a circular arrays for the space-time adaptive processing which is applicable to the airborne-radar, although this Uniform Linear Array (ULA) can be found applicable in radar. The airborne-radar scenario has been summarized and completed by Sarkar and Adve in ref. [3]. A pulsed Doppler radar was considered which consist of a circular phased array positioned on airborne board moving at a continual velocity. Wideband signals was still not achieved.

In this research work, wideband beamforming issue has been considered. This is one of the fundamental functions of smart antenna system. Predominantly, the recommended techniques in the literature are narrow-band beamforming. The bandwidth ratio to center frequency of the narrowband incoming signal is ≤ 1%. Wideband applications are recommended in the work for fractional bandwidth up to 100-150% [16]. In this present research, Uniform Linear Arrays (ULA) has been used for both method. Measurements of spatial frequencies is equivalent to direction finding. Specifically, a main lobe must be generated by the array in the direction of a desired incoming signals called signal-of-interest (SOI) and numerous nulls towards respective undesired or interference incoming signals [19]-[23].

The paper is organized as follows. Section II describes the early forms of spatial signal processing, while the Section III describes the preliminaries of array signal processing model for the antenna array receiving signal arbitrarily, and a linear antenna using omni-directional antenna elements (receiving a narrow-band signal). The wideband beamforming technique has been presented in the Section IV followed by the various spatial processing techniques for the antenna arrays in Section V. Section VI describes the evaluation of direction of arrival (DOA). Finally, Section VII concludes the work and recommend the future aspects.

II. Preliminaries of Signal Processing Array Model

Let us assume a linear antenna array of *N* units having an omni-directional antenna. The received signal by this narrowband antenna element array at any instant time, *k* can be stated mathematically as:

$$z_n(k) = x_n(k) + j(k) + q_n(k) \quad (1)$$
$$= (k)\left[x_n + j + q_n\right]$$

where $x_n(k)$ is the *Nx1* vector of the signal, $j(k)$ denotes the *Nx1* interference signal's vector, and $q_n(k)$ represents the *Nx1* vector of the noise output. Assuming that the desired signal, interference and noise are statistically independent, the desired signal can be expressed as:

$$\mathbf{Z_n}(\mathrm{k}) = \mathrm{Z_n}(\mathrm{k})\mathbf{a}_\mathrm{s} \quad (2)$$

where $z(k)$ is the preferred received signal waveform. $\boldsymbol{a}_s$ is the steering vector connected to the spatial signature and if substituted into equation (1), it becomes the oriented vector. The beamformer output for the narrowband can be expressed as:

$$\mathrm{z(k)} = \mathbf{K}^{\mathrm{H}}\mathbf{N}(\mathrm{k}) \quad (3)$$

The beamforming algorithm for the weight vector (K) is $(k_1, k_2, \ldots, k_x)^T$

The *Nx1* beamformer complex weight vector of the antenna array is *k*, while $(.)^{\boldsymbol{H}}$ is the Hermitian transpose, and $n(t)$ is the Nx1 array shot vector.

The signal to interference plus noise ratio (SINR) for the defined beamforming problem can be written as:

$$SINR = \frac{P}{N + I} \quad (4)$$

where *P, N and I* are the array output signal power, noise power output of the array signal, and the interference power, respectively.

The fundamental issue is the interference (*I*), taken at a point source $x \in \mathbb{R}^k$ expressed as [20]

$$I(x) = \sum_{y \in \tau} p_y h_y \ell(\|x - y\|) \quad (5)$$

where $\tau \subset \Re^k \Rightarrow$ set of all receiving signals, p_y is the power received by the signal *y*, h_y is the fading coefficient of the signal power, and ℓ the path loss function. We assumed $\|x - y\|$ to depend on the distance from the antenna element y to another point of antenna element x. Therefore,

$$SNR = \frac{w^H R_s w}{w^H R_{i+n} w + I} \quad (6)$$

Hence, SNR

$$= \frac{w^H E\left\{x(k)s^H(k)\right\}w}{w^H E\left\{\varepsilon\right\}w + I(x)} \quad (7)$$

ε in Eq. (7) is $(j(k)+n(k))(j(k)+n(k))^H$. Hence, SNR

$$=\frac{w^H E\{x(k)s^H(k)\}w}{w^H E\{(j(k)+n(k))(j(k)+n(k))^H\}w+\sum_{y\in\tau} p_y h_y \ell(\|x-y\|)} \quad (8)$$

$E\{x(k)x^H(k)\}$, and $E(\varepsilon)$ are the desired signal, while $\sum_{y\in\tau} p_y h_y \ell(\|x-y\|)$, is the fundamental issue of the interference power taken into consideration for this work, respectively, $E\{\bullet\}$ is the statistical expectation. Considering an extensive wireless system, τ, h_y, and P_y are the unknowns.

The interfering signal points and the path loss can control the interference to the first order.

The covariance matrix of the corresponding desired signal in Eq. (6), can be of an arbitrary rank as shown in reference [25], i.e. $1\le rank\{R_s\}\le M$. $rank\{R_s\}>1$ for signals that have randomly fluctuating wavefronts. This might occur frequently in wireless communications. For a point source signal, $R_s = 1$. From Eq. (6), R_{i+n} known as interference plus noise covariance matrix and can be substituted using the covariance matrix for the data sample as in ref. [25].

$$\hat{Z}=\frac{1}{n}\sum_{i=1}^{n} y_i(x)y_i^H(x) \quad (9)$$

n in Eq. (9) is the number of samples of training data

III. Array Signal Processing Model

Consider a plane wave signal for random antenna array as in Fig. 1 receives signal in the direction $-\hat{s}_r(\theta,\phi)$, the analytical signal of the plane wave signal can be expressed as:

$$x(t)=y(t)e^{j\omega t} \quad (10)$$

where $x(t)$ is the signal along the plane wave, $y(t)$ is the signal along the base-band or the real signal, and ω is the frequency carrier, and t is the measured time respectively. q1, q2, qi, and qm are the antenna elements in Fig. 1.

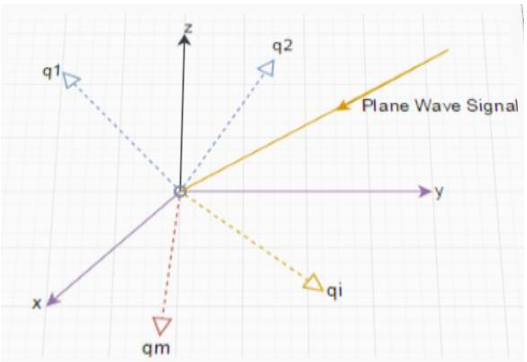


Fig. 1. The system coordinate signal for the random antenna array.

Taking the real part of the analytical signal, the real signal r(t) can be expressed as:

$$r(t)=\mathrm{Re}\{x(t)\}=y(t)\cos(\omega t) \quad (11)$$

The mathematical correlation between the Phasor-form signal b(t) and the analytical signal a(t) can be written as:

$$x(t)=Xe^{jwt} \Rightarrow X=y(t) \quad (12)$$

The power of r(t) is

$$r(t)=\frac{1}{2}|z(t)|^2$$

The phasor-form signal for the r(t) can be expressed as:

$$\frac{1}{2}E\{s(t)s*(t)\}=\frac{1}{2}SS*$$

The Phasor-form signal "s is a function of (x, y, z) and t" and also a complex quantity. Mathematically,
S = s(x, y, z) and s = s(t).

Characterization of antenna element's time domain can be normalized by the impulse response. Fast transient pulses are responsible for antenna-excitation in time domain systems. Firstly, we have considered the impulse and frequency response of the system. The impulse response and $k(t)$ relates the antenna elements to the plane wave signal $z(t)$. The input to the system is $z(t)$, while the antenna elements act as the output $q(t)$. From the convolution integral,

$$q(t)=\int_{-\infty}^{+\infty} k(\tau)z(t-\tau)d\tau \quad (13)$$

Using the convolution operator $\otimes$, (13) can be written as:

$$q(t)=k(t)\otimes z(t)$$

If the system is excited by $\delta(t)$, i.e. the Dirac impulse, Eq. (14) and (15) is derived

$$q(t)=k(t)\otimes z(t)=k(t) \quad (14)$$

$$q_i(t)=f_i(\theta,\phi)\delta\kappa \quad (15)$$

$f_i(\theta,\phi)$ is the radiation pattern for the ith antenna element, δk is the Dirac impulse response changes with $\kappa=(t-\tau_i)$, where

$$\tau_i=\frac{q_i\cdot(-\hat{s}_r(\theta,\phi))}{v} \quad (16)$$

and v is the propagation speed. Hence, the impulse response (IR) of the antenna elements in relation to the signal plane wave z(t) is:

$$q_i(t)=f_i(\theta,\phi)\delta\left[t-\frac{q_i\cdot(-\hat{s}_r(\theta,\phi))}{v}\right] \quad (17)$$

i = 1, 2, ... m, where m is the total number of antenna elements in the array and t is the reference time of the coordinate system. If the radiation patterns of the system is isotropic, $f_i(\theta,\phi)$ will be equal to one, and

$$q_i(t)=\delta\left[t-\frac{q_i\cdot(-\hat{s}_r(\theta,\phi))}{v}\right]=\delta(t-\tau_i) \quad (18)$$

Signal received at the *ith* elements will be:

$$z_i(t) = k_i(t) * s(t) + n_i(t) \tag{19}$$

$$= \delta\left[t - \frac{q_i \cdot (-\hat{s}_r(\theta,\phi))}{v}\right] * s(t)e^{j\omega t} + n_i(t)$$

$$= s(t-\tau_i)e^{j\omega t}e^{jkp_i.\hat{a}_r(\theta,\phi)} + n_i(t)$$

$$\approx z(t)e^{jkq_i.\hat{s}_r(\theta,\phi)} + n_i(t)$$

$$= s(t) + n_i(t) \tag{23}$$

where $n_i(t)$ is the noise in the antenna array elements. Hence, the array signal vector received by the antenna elements is:

$$\mathbf{q} = \mathbf{s} + \mathbf{n}$$
$$= s(t)\mathbf{b} + \mathbf{n} \tag{24}$$

where

$$\mathbf{q} = \begin{bmatrix} q_1(t) \\ q_2(t) \\ \vdots \\ q_N(t) \end{bmatrix}, \mathbf{b} = \begin{bmatrix} e^{jkq_1.\hat{s}_r(\theta,\phi)} \\ e^{jkq_2.\hat{s}_r(\theta,\phi)} \\ \vdots \\ e^{jkq_m.\hat{s}_r(\theta,\phi)} \end{bmatrix}, \mathbf{n} = \begin{bmatrix} n_1(t) \\ n_2(t) \\ \vdots \\ n_N(t) \end{bmatrix} \tag{25}$$

If there are sets of signals k $x_j(t) = s_j(t)e^{j\omega t}$ entering the system in the direction (θ_j, ϕ_j), $j = 1, 2, \ldots, k$, the array signals received is the array signal vector received by the antenna elements in Eq. (24):

$$\boldsymbol{q} = \boldsymbol{s} + \boldsymbol{n}$$
$$= s_1(t)\boldsymbol{b}_1 + s_2(t)\boldsymbol{b}_2 + \cdots + a_k(t)\boldsymbol{b}_k + \boldsymbol{n}$$
$$= [b_1, b_2, \ldots b_k][s_1(t), s_2(t), \ldots s_k(t)]^T + n$$
$$= \boldsymbol{B}_a + \boldsymbol{n} \tag{26}$$

where $B = [b_1, b_2, \cdots, b_k]$

$$= \begin{bmatrix} e^{jkq_1.\hat{s}_r(\theta_1,\phi_1)} & e^{jkq_1.\hat{s}_r(\theta_2,\phi_2)} & \cdots & e^{jkq_1.\hat{s}_r(\theta_k,\phi_k)} \\ e^{jkq_2.\hat{s}_r(\theta_1,\phi_1)} & e^{jkq_2.\hat{s}_r(\theta_2,\phi_2)} & \cdots & e^{jkq_2.\hat{s}_r(\theta_k,\phi_k)} \\ \vdots & \vdots & \ddots & \vdots \\ e^{jkq_m.\hat{s}_r(\theta_1,\phi_1)} & e^{jkq_m.\hat{s}_r(\theta_2,\phi_2)} & \cdots & e^{jkq_m.\hat{s}_r(\theta_k,\phi_k)} \end{bmatrix}$$

$$s = \begin{bmatrix} s_1(t) \\ s_2(t) \\ \vdots \\ s_k(t) \end{bmatrix}$$

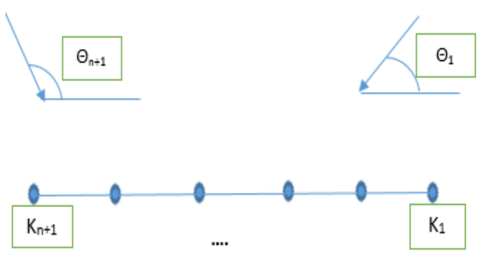


Fig. 2. Uniform linear array inter-element spacing.

Fig. 2 shows a uniform linear antenna array having an equal inter-element spacing *d*, taken the first element from the origin, we can have the matrix **B** in the form of:

$B = [b_1, b_2, \ldots, b_k]$

$$= \begin{bmatrix} 1 & 1 & \cdots & 1 \\ e^{jkd\sin\theta_1\cos\phi_1} & e^{jkd\sin\theta_2\cos\phi_2} & \cdots & e^{jkd\sin\theta_k\cos\phi_k} \\ \vdots & \vdots & \ddots & \vdots \\ e^{jk(2m-1)d\sin\theta_1\cos\phi_1} & e^{jk(2m-1)d\sin\theta_2\cos\phi_2} & \cdots & e^{jk(2m-1)d\sin\theta_k\cos\phi_k} \end{bmatrix} \tag{27}$$

The average array gain when the signal direction is uniform at [-0.1 0.1] is studied in relation to the SNR in Eq. (6) of section II. The result is shown in Fig. 3.

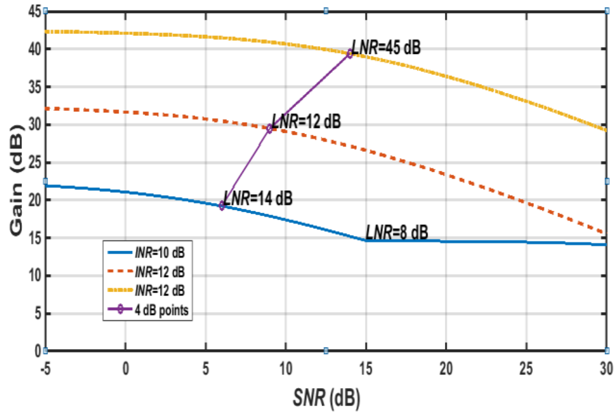


Fig. 3. Gain versus SNR for the antenna element

The total antenna elements considered for the simulation of the array gain is 16, while the sigma range [-0.1:1/1000:0.1]. For the 4 dB points observed in the simulation, when INR = 10 dB, it has an optimum gain of 18 dB and SNR of 6 dB, while for INR of 12 dB, its optimum gain is 29 dB while its SNR is 8 dB, and for INR of 12 dB, optimum gain is 39dB and SNR is 14 dB.

IV. Beamforming Technique

One of the utmost significant techniques in smart antennas, is beamforming [1]. Beamforming/spatial filtering is a signal processing technique that finds a wide application in smart antenna systems. The antenna can produce one or more beams pattern towards the signal of interest (SOI) paths/angle, and simultaneously nulls in the signal direction of no interest (SNOI). If the direction of the antenna remains constant on the received signal, there will be loss of signal [1], [21]. Beamforming technique can generally be categorized into fixed and adaptive. During the operation of fixed beamformer, parameters are fixed whereas in adaptive beamformer, parameters are updated constantly based on the signals received.

A. *Wideband Beamforming*

For wideband beamforming, two techniques are popularly used for the signal processing analysis.

(i) time-domain (TD) processing, and

(ii) frequency-domain (FD) processing [16].

These methods can generate frequency invariant beam patterns for wideband signal. One of the advantages of frequency-domain approach over the time-domain for signals with large bandwidths is the computational approach. Fig. 4 shows the simulation of a frequency invariant beam-pattern in the normalized frequency band of fractional bandwidth of 120%. The frequencies are normalized over the sampling frequency.

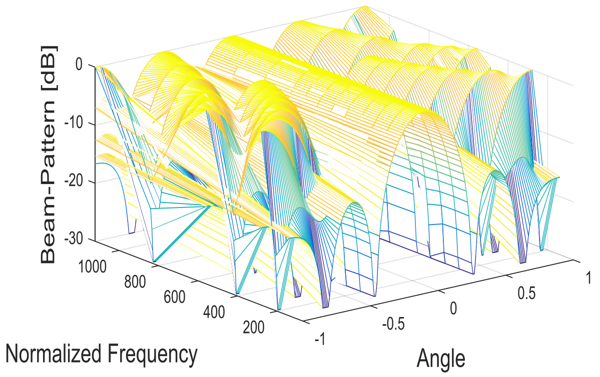


Fig. 4. Frequency invariant wideband beamforming pattern

Some of the advantages of frequency invariant wideband beamforming are: (i) faster convergence speed and (ii) lower computational complexity. The effective approach to solve the blind wideband beamforming problem is the frequency invariant beamforming technique.

B. *The Optimum Beamforming*

In optimum beamforming, the knowledge of the desired signal characteristic is required. This can be based on either its statistics of:

(i) Maximum SNR, and

(ii) Reference signal method.

If the knowledge of the desired signal required is inaccurate, the desired signal will be attenuated by the optimum beamformer as if it were interference [21]. For optimum beamformer array, the outputs sensor are joined by a weight vector in order to receive desired signal without distortion, while simultaneously rejecting interfering signals to the barest minimum [26]-[27].

A satisfactory quality of service at economical rate is always delivered by the optimum beamforming system. This continually serve numerous users. Whenever the broadband signals is of reference point, beamforming can be implemented in the frequency domain.

C. *Beampattern for Broadband Signal*

The source of an antenna array system can be modelled in a wideband manner and potentially non-stationary random process. Assuming a discrete aperture of an array of N sensors in which the wave field is sampled in space and time. Assumed in the direction of θ_0, the source lies on the plane of the array, and plane wave impinged signal on the array. For a broadband signal (for the plane wave), response to this signal from the plane wave will characterize the beampattern for broadband signal of the array. Therefore, *nth* sensor output modelled at time t is [28]:

$$x_n(k) = \delta[k - \tau_n(\theta_o)], \tag{28}$$

where $n = 1, 2, \ldots, N$, $x_n(k)$ is the output sensor n at time k, and $\delta(k)$ is the corresponding of $n = 1, 2, \ldots, N$ signal that processes distinguishing source signal, $\tau_n(\theta_0)$ is the time propagation from the signal source to sensor n, while θ_0 is the direction of signal of the source. *Dmochowski "et al."* [28] indicated that noise is had not been put into consideration, but the noise is considered in this present work. Since the direction of incoming signal is known as θ_0, the weights are chosen to maximize the array output of the SNR.

The output noise power and output signal power for the array is mathematically expressed as:

$$\begin{aligned} p &= E\{y_x^* y_x\} = w^H E\{x^* x^T\} w \\ N &= E\{y_n^* y_n\} = w^H E\{nn^H\} w = \sigma^2 w^H I_N w = \sigma^2 w^H w \\ &= \sigma^2 \sum_{i=1}^{N} |w_i|^2 \end{aligned} \tag{29}$$

Therefore, SNR for the sensors will be

$$SNR = \frac{P}{N} = \frac{w^H E\{x^* x^T\} w}{\sigma^2 \sum_{i=1}^{N} |w_i|^2} \tag{30}$$

From (29)

$$\begin{aligned} & w^H E\{x^* x^T\} w \\ &= w^H E\{x^*(k) x^* x(k) x^T\} w \\ &= p w^H x^x x^T w \\ &= p \left(\sum_{j=1}^{N} w_j z_j\right)^* \left(\sum_{j=1}^{N} w_j z_j\right) \\ &= \left|\sum_{j=1}^{N} w_j z_j\right|^2 \\ &\le p \sum_{j=1}^{N} |w_j|^2 \sum_{j=1}^{N} |z_j|^2 \end{aligned} \tag{31}$$

Eq. (31) is achieved using Schwarz inequality.

Hence, ($p = E\{s^\cdot(t)s(t)\}$ where p is the baseband signal power. Therefore,

$$SNR = \frac{w^H E\{s^* s^T\} w}{\delta^2 \sum_{j=1}^{N} |w_j|^2} \le \frac{p \sum_{j=1}^{N} |w_j|^2 \sum_{j=1}^{N} |z_j|^2}{\delta^2 \sum_{j=1}^{N} |w_j|^2} = \frac{p}{\delta^2} \sum_{j=1}^{N} |z_j|^2 \tag{32}$$

Simulated results for the array gain and SNR are shown in Fig. 5.

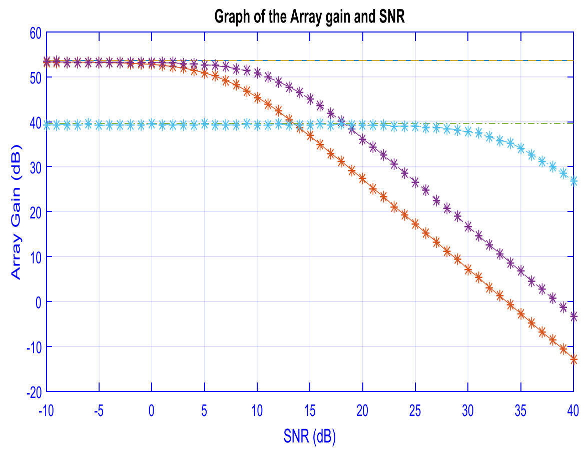


Fig. 5. Simulated results for the array gain and SNR.

From Eq. (31)

$$p = w^H E\left\{x^* x^T\right\} w$$

$$= w^H E\left\{x x^T\right\}^* w$$

$$= w^H R_{xx}^* w \qquad (33)$$

where

$$R_{xx} = E\left\{xx^H\right\} = \begin{bmatrix} E\{x_1x_1^*\} & E\{x_1x_2^*\} & \cdots & E\{x_1x_N^*\} \\ E\{x_2x_1^*\} & E\{x_2x_2^*\} & \cdots & E\{x_2x_N^*\} \\ \vdots & \vdots & \ddots & \vdots \\ E\{x_Nx_1^*\} & E\{x_Nx_1^*\} & \cdots & E\{x_2x_N^*\} \end{bmatrix} \qquad (34)$$

R_{xx} is the data correlation matrix if the average values of $x_1(t), x_1(t), ..., x_N(t)$ =0. The data correlation matrix can be used to analyze wave propagation.

In Section III, an *M* equally spaced of linear antenna array units of omni-directional antenna was considered. Assuming the inter-element spacing along the x-direction is *d*. The spatial aliasing exist if *x* is relative to *λ*.

Assuming the plane wave for a uniform linear array is changed from the plane wave considered before to a monochromatic plane, its response delay-and-sum beamformer tuned to $\vec{\alpha}$ can be expressed as:

$$z(t) = w(\omega^o\vec{\alpha} - \vec{k}^o)\ell^{j\omega^o t} \qquad (35)$$

$$w(\vec{k}) = \frac{\sin\left(mk_x \frac{d}{2}\right)}{\sin\left(k_x \frac{d}{2}\right)}$$

$$w(\omega^o\vec{\alpha} - \vec{k}^o) = \frac{\sin\left(m[\omega^o\alpha_x - k_x^o]\frac{d}{2}\right)}{\sin\left([\omega^o\alpha_x - k_x^o]\frac{d}{2}\right)}$$

Using $k_x = \omega^o\alpha_x$

$$W(\omega^o\vec{\alpha} - \vec{k}^o) = \frac{\sin\left(M\left[k_x - k_x^o\right]\frac{d}{2}\right)}{\sin\left(\left[k_x - k_x^o\right]\frac{d}{2}\right)} \qquad (36)$$

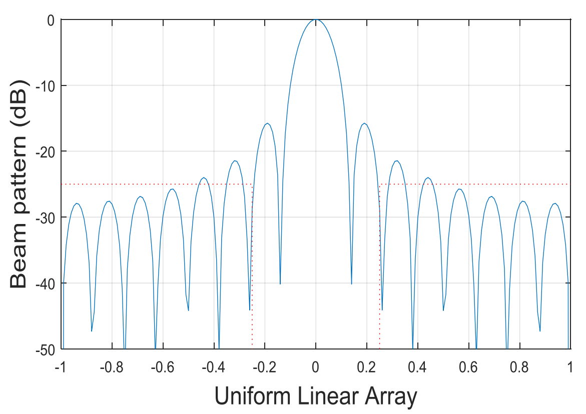
Let us consider it in terms of angles, let

$$k_x = \left(\frac{2\pi}{\delta}\right)\sin(\phi)$$

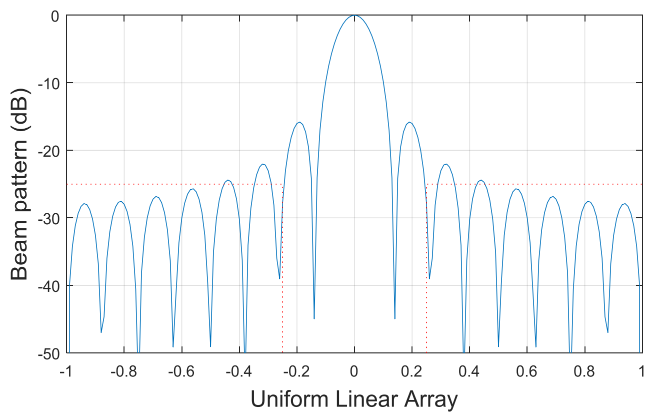
$$W(k_x - \vec{k}^o) = \frac{\sin\left(M\frac{\pi}{\delta}\left[\sin\phi^o - \sin\phi\right]d\right)}{\sin\left(\frac{\pi}{\delta}\left[\sin\phi^o - \sin\phi\right]d\right)} \qquad (37)$$

The beam patterns for the uniform array has been illustrated in Fig. 6 (a-f).

A beamformer can be steered to a direction of arrival *θ*, if a complex weight is attached to each sensor and then take the addition through the aperture.



(a)



(b)

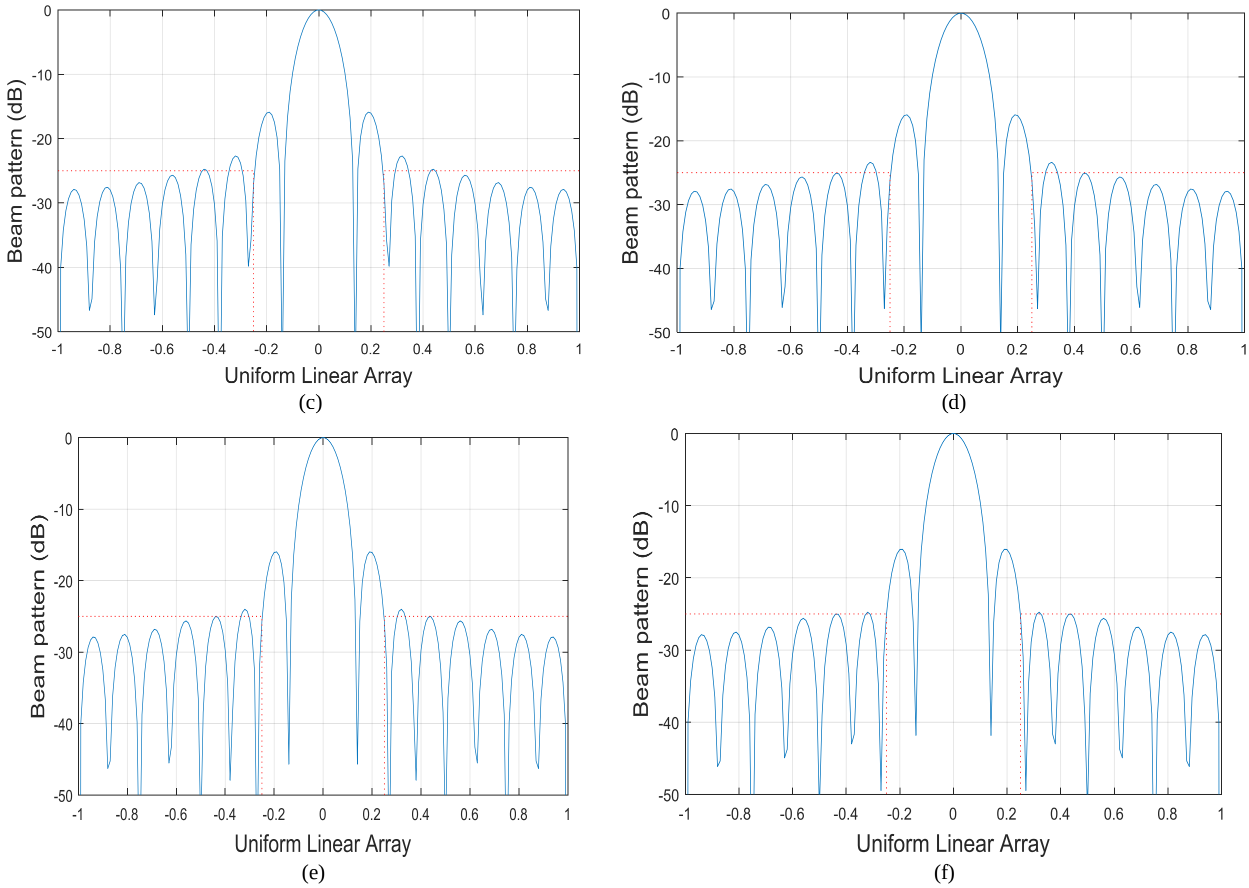


Fig. 6. Beam patterns and the uniform linear array

V. Spatial Techniques of Antenna Arrays

A smart antenna is spatially sensitive and has inherent intelligence to create a beam with high gain in the preferred direction. It can change the beam pattern actively and can find and track the users as necessary [20].

Higher degrees of freedom can be achieved in smart antenna design system using a spatial processing, which will improve the overall performance of the system [9]. Smart antenna can separate signals from various users who are separated in space (i.e. by distance), but utilizing the same radio channel (i.e. center frequency, time-slot, and/or code); this procedure is called space division multiple access (SDMA).

Various signal processing applications are employed for estimating some parameters or the whole waveform of the received signals [21].

A. Spatial Smoothing Technique

Spatial smoothing processing of antenna arrays is one of the conventional methods for improving Multiple Signal Classification (MUSIC) algorithm. This method makes the algorithm to work even in existence of coherent signals. Among the recommended methods to decorrelate coherent signals, spatial smoothing is one of the effective techniques, widely use in wireless communication systems. This technique is established on a preprogramming structure that distributes the entire array into corresponding subarrays, and then find the arithmetic mean of the subarrays output covariance matrices. This resulted in spatially smoothed covariance matrix [29]-[31].

B. Spatial Filtering

Spatial sampling in one dimension which can be written as:

$$\begin{aligned} Z(\omega\vec{\delta}_j^o,\omega) &= \sum\nolimits_i S_i(\omega) w(\omega\vec{\delta}_j^o - \omega\vec{\delta}_i^o) \\ &= S_j(\omega) w(0) + \sum\nolimits_{i\neq j} S_i(\omega) w(\omega[\vec{\delta}_j^o - \vec{\delta}_i^o]) \end{aligned} \quad (38)$$

If we can design w so that we have a spatial filter for direction α_j

$$W(\omega[\vec{\alpha}_j^\circ - \vec{\alpha}_i^\circ])\Delta W(0) \quad (39)$$

for $i \neq j$

Aperture is a spatial region that transmits or receives propagating waves. For space-time field through aperture:

$$z(\vec{x},k) = \omega(\vec{x}) f(\vec{x},k) \quad (40)$$

Spatial domain multiplication is a convolution in wavenumber domain

$$z(\vec{k},\omega) = \frac{1}{(2\pi)^3}\int_{-\infty}^{\infty} W(\vec{k}-\vec{l}) F(\vec{l},\omega) d\vec{l} \quad (41)$$

where $W(\vec{k}-\vec{l})$ is the spatial FT of w, $F(\vec{l},\omega)$ is the spatiotemporal FT of f,

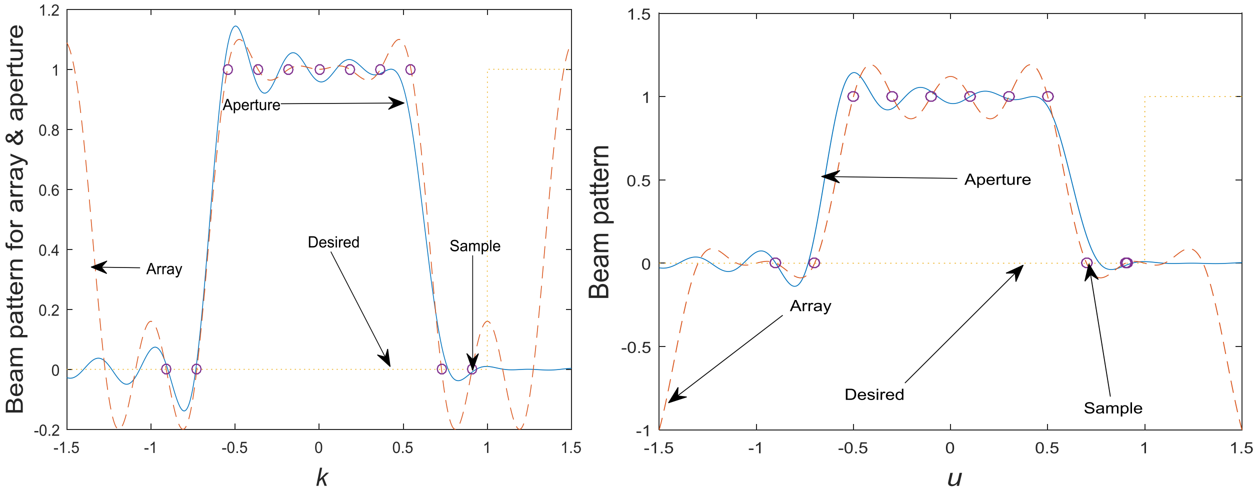
$$W(\vec{k}) = \int_{-\infty}^{\infty} w(\vec{k}) \exp\left\{\oplus j\vec{k}.\vec{x}\right\} d\vec{x}$$

$W(\vec{k})$ =aperture smoothing and $\oplus$ =Mathematician's FT For plane wave,

$$\begin{aligned} f(\vec{x},t) &= s(t-\vec{\alpha}^{\circ}\cdot\vec{x}) \\ F(\vec{k},\omega) &= S(\omega)(2\pi)^3\delta(\vec{k}-\omega\vec{\alpha}^{\circ}) \end{aligned} \quad (42)$$

Smoothed in wavenumber space:

$$\begin{aligned} Z(\vec{k},\omega) &= \frac{1}{(2\pi)^3}(W * \vec{x}F)(\vec{k},\omega) \\ &= s(\omega)W(\vec{k}-\omega\vec{\alpha}^{\circ}) \end{aligned} \quad (43)$$



(a) (b)

Fig. 7. Beam pattern array aperture and their corresponding angle

C. Delay-and-Sum Beamforming Technique

Delay-and-sum beamforming method is a conventional methods for DOA estimation of array of antenna signals also known as classical beamformer, having equal magnitudes in weights [2]. The selected array phases are steered in a specified direction, called the look direction. In the look direction, the source power is equivalent to the mean output power of classical beamformer driven in the look direction.

Consider a sensors of M positioning at $\vec{x}_{\circ}\ldots\vec{x}_{M-1}$. If the phase center is put at the origin, it can be expressed as:

$$\sum_{m=0}^{M-1}\vec{x}_m = \vec{0} \quad (44)$$

Its delay-and-sum beamforming can be found as

$$z(t) = \sum_{m=0}^{M-1} w_m y_m(t-\Delta_m) \quad (45)$$

Let us consider the Monochromatic plane waves,

$$\begin{aligned} f(\vec{x},k) &= \exp\left\{j\omega^0(k-\vec{\alpha}^{o}.\vec{x})\right\} \quad (46) \\ &= s(k-\alpha^{o}\cdot\vec{x}) \end{aligned}$$

where $s(k) = \exp(j\omega^{o}k)$

For a plane wave, the delay-and-sum beamformer response is

$$\begin{aligned} z(t) &= \sum_{m=o}^{M-1} w_m s(k+(\vec{\alpha}-\vec{\alpha}^{o})\cdot\vec{x}_m) \quad (47) \\ &= \sum_{m=o}^{M-1} w_m \exp(j\omega^{o}[k+(\vec{\alpha}-\vec{\alpha}^{o})\cdot\vec{x}_m]) \\ &= \left[\sum_{m=o}^{M-1} w_m \exp(j\omega^{o}(\vec{\alpha}-\vec{\alpha}^{o})\cdot\vec{x}_m)\right]\exp(j\omega^{o}t) \\ \vec{k}^{o} &= \omega^{o}\alpha^{o} \\ &= \left[\sum_{m=o}^{M-1} w_m \exp(j(\omega^{o}\vec{\alpha}-\vec{k}^{o})\cdot\vec{x}_m)\right]\exp(j\omega^{o}t) \\ &= W(\omega^{o}\vec{\alpha}-\vec{k}^{o})\exp(j\omega^{o}t) \end{aligned}$$

where the discrete aperture smoothing function is

$$W(\vec{k}) = \sum_{m=o}^{M-1} w_m \exp(j\vec{k}.\vec{x}_m)$$

D. Deterministic Beamformer

Adaptive algorithms proposed for beamforming are the Howells-Applebaum adaptive loop [19, 20]. Suppose we have an array consisting of M as odd elements evenly spaced along the x-axis, centered at the origin, with spacing d. The twist here is that each element, instead of being a point, is actually a small linear segment of length L along the x-axis (where $L<d$, so the segments don't overlap), centered at the element locations. The wave number-frequency response of the resulting filter and sum beamformer can be found if the sensor delays are adjusted to steer the beam to look for plane waves propagating with a slowness vector $\vec{\alpha}$.

We have observed a space-time field *m'th* sensor position, $\vec{x}_m$. $f(\vec{x}_m, k)$, our sensors can only gather energy over a finite area, indicated by the (spatial) aperture function $\omega(\vec{x})$. $f(\vec{x}_m, k)$ is the field values. Therefore,

$$\omega(\vec{x}) = \begin{cases} \neq 0 & inside \quad aperture \\ = 0 & outside \quad aperture \end{cases} \tag{48}$$

where $y_m(t)$ is the sensors output. If sensor is perfect, $y_m = \kappa . f(x_m, k)$

Directional sensors have significant spatial extent. They spatially integrate energy over the aperture, i.e. they focus a particular propagation direction. e.g. parabolic dish. They are described by the aperture function, $\omega(\vec{x})$, which describes: (i) spatial extent reflects size and shape, (ii) aperture weighting: relative weighting of the field within the aperture (also known as shading, tapering, apodization). For places where $\omega(\vec{x}) \neq 0$, we sometimes get to pick $\omega(\vec{x})$ called aperture weighting, or apodization, shading, or tapering.

VI. Direction of Arrival (DOA) Approximation

The direction of arrival approximation is one of the foremost applications of linear arrays [1], [31]-[33]. As mentioned earlier in this work, some optimum beamformers need the DOA information to control the optimum weight vector. Various research work had been investigated and carried out on DOA estimation algorithms which are fundamentally categorized into two, spectral based and parametric technique. Some of the spectral based technique uses Bartlett method. These includes, Capon's Minimum Variance Method and Eigen Structure Method (Spectral MUSIC) [30], [31]. In terms of accuracy and resolution, parametric method yield a better performance. The cost of the performance increase of parametric method gives a higher complexity and this will require more computation. The most attractive parametric methods are the Root MUSIC and ESPIRIT.

Smart antennas performance and optimization depends on the idea that the DOA of signal of interest is recognized to the system by means of one of the techniques classified in ref. [4]. Better arrays estimation capacity can be achieved by taking the DOA distribution into deliberation [31], [32].

The Direction of arrival of signals must match the phase delay variances of signals at the outputs of receiving antenna array elements to avoid loss of signals. The level of uncertainty in determining DOA depends on the geometry of array, parameters, and characteristics of the array elements (sensors) [32]-[35].

There are several techniques of evaluating the direction/angle of arrival estimate in smart antenna. These includes: (i) mechanical or electronic beampattern antennas controlled reception, (ii) interferometry phase method, subspace-based. Some of the techniques that are commonly used in solving the direction of arrival in smart antenna are MUSIC, ESPRIT, etc. This can be found in ref. [29]-[31].

The MUSIC for performing DOA estimation because it does not take any advantage of the array geometrical configuration. The array of interest is assumed to consist of *N* isotropic elements that receive directional signals from the far field. All propagating signals are modeled as narrowband. The direction of arrival (DOA) of a signal is mainly calculated by using sensors or arrays of antenna [31]. The DOA estimation has been carried out in ref. [33]-[37], and the process were described in detail for uncorrelated signal sources. In Fig. 2, section III of this present work, the direction of signal from the received signals can be estimated using the following description: the set of incoming signal direction is θ_1, the elements *N* of the antenna array are in a linear equispaced. Here, the number of incoming signals are unknown, likewise the direction and the amplitude. The signals are normally corrupted by noise. Let the number of the unknown signals be *M*, *M*<*N*, and with the assumption of white Gaussian noise,

$$x = v(\theta) + n \tag{49}$$

where *x* is the length *N* vector of the received signals, $\theta = \left[\theta_1, \cdots, \theta_p\right]^T$ is the set of parameters, and ***v*** is a known function of parameters, $\text{var}(\theta_p) \geq J_{pp}^{-1}$

$$J_{ij} = E\left\{\frac{\partial^2}{\partial_{\theta i}\partial_{\theta j}}[\ln fx(x/\theta]\right\}$$

J is known as the Fischer Information Matrix. Single signal corrupted by noise can be expressed as:

$$x = \alpha s(\theta) + n \tag{50}$$

where **s** is the direction-finding vector of the signal, **n** is zero mean Gaussian with covariance $\delta^2 I$

Assuming $\quad \alpha = ae^{jb}$

$$\theta = [a, b, \theta]^T$$

In this situation,

$$v(\theta) = \alpha s(\delta)$$

$$fx(x/\phi) = Ce^{-(x-v)^H R^{-1}(x-v)}$$

$$\ln fx(x/\theta) = \ln C + \frac{-x^H x + \alpha^* s^H(\phi) x + \alpha x^H s(\phi) - |\alpha|^2 s^H(\phi) s(\phi)}{\sigma^2} \tag{51}$$

$$g(\theta) = \frac{1}{\sigma^2}\left[ae^{-jb} s^H(\phi) x + ae^{jb} x^H s(\phi) - a^2 s^H(\phi) s(\phi)\right] \tag{52}$$

Therefore, CRB estimation problem for the DOA is:

$$\text{var}(\theta) \geq \left[E\left(\frac{\partial^2 g}{\partial \theta^2}\right)\right]^{-1} \tag{53}$$

$$\geq \frac{6\delta^2}{|\alpha|^2 N(N^2-1)(kd)^2 \sin^2\phi}$$

A. Uncorrelated Signal Sources Estimation

The DOAs estimation of the uncorrelated signal sources must be estimated in the first instance. Let us consider a group of coherent sources equivalent to a virtual source. The Eigen decomposition can be expressed as [25]

$$R_s = U\Sigma U^H = U_s\Sigma_s U_s^H + U_n\Sigma_n U_n^H \tag{54}$$

Assuming $R_s = Q \wedge Q^H$, then $R = Q[\wedge + \sigma^2 I]Q^H$

Therefore,

$$R = QQ^H \begin{bmatrix} \lambda_m+\sigma^2 & 0 & \cdots & 0 & 0 & \cdots & 0 \\ 0 & \lambda_m+\sigma^2 & \cdots & 0 & 0 & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \cdots & \lambda_m^2+\sigma^2 & 0 & \cdots & 0 \\ 0 & 0 & \cdots & 0 & \sigma^2 & \cdots & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & 0 & 0 & \cdots & \sigma^2 \end{bmatrix} \tag{55}$$

The Eigen vector matrix (Q) can be partitioned into noise subspace (Q_n) and signal matrix (Q_s), columns M of Q_s » signal Eigen values of M, N-M columns of Q_n » noise Eigen values

The m-th signal Eigen value is written as:

$$\lambda_m + \sigma^2 = N|\alpha_m|^2 + \sigma^2 \tag{56}$$

By orthogonality, Q and Q_s are ⊥ to Q_n

Hence, noise Eigen vectors are ⊥ to the steering signal vectors. The Eigen values for the uncorrelated sources are shown in Fig. 8.

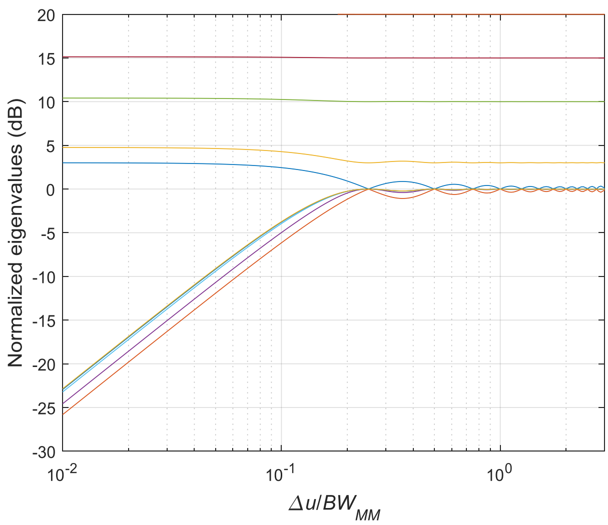


Fig. 8. Eigen values for uncorrelated sources.

An array of ten omnidirectional sensors/antennas that involves three linear subarrays was used. The first subarray involves of four sensors that has an inter-element spacing of: (i) 0.42λ, (ii) 0.53λ, and (iii) 0.32λ, where λ is the signal wavelength. The plot shown indicates that the beamformer can be steered to the desired direction of the main beam. The plot also shown the simulation results when the interference signals are coming from different directions.

VII. Conclusions and Future Aspects

In this work, we have analyzed smart antenna using spatial signal processing. Using a spatial signal processing, this will improve the SNR performances of uplink and downlink. With beamforming technique, the signal/noise level, speed, and terminal location has been looked on. The defined problem for signal to interference plus noise ratio (SNR) established in section II, which has been neglected by researchers for not taking interference as the fundamental issue. Wideband beamforming issue has also been considered. Most of the techniques used in the literature are narrowband beamforming.

As a future work, the optimized smart antenna array can be fabricated and its performance can be evaluated using the improved signal quality and spatial processing. The proposed method can be used to determine the smart antenna parameters. More degree of freedom can be achieved in the system design using spatial processing, which can also improves the overall performance of the system.

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