Tightly Coupled Dipole Array with Integrated Phase Shifters for Millimeter-Wave Connectivity

DISsertation

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By

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Abstract

Advanced satellite communications (SATCOM) applications and multifunctional apertures at Ka-band have created a growing interest for compact, agile, beam scanning, wideband antennas. Phased arrays are of particular interest as they offer beam-steering agility, lower profile and much wider bandwidth compared to conventional mechanical beam steering techniques that are bulky, introduce mechanical noise, and quite slow in steering the beam. Recently introduced tightly coupled dipole arrays (TCDAs) achieve ultra-wide bandwidth performance while having extremely low profiles. However, the size, weight, loss, and complexity of the feed network still make such front ends impractical for many applications. In addition, scaling the TCDA to a higher frequency bands, such as Ka-band, creates issues that have never been addressed before. Most importantly, fabrication limitations arise due to size reduction and material loss at increased frequencies must be addressed in the design. The development of low-cost, simple, lightweight and low-loss wideband phased arrays at Ka-band is critical to improve the capabilities of small platforms such as wireless sensors and unmanned aerial vehicles.

This work represents a novel TCDA at Ka-band and addresses the following challenges: 1) An accurate equivalent circuit for the TCDA unit cell at Ka-band; 2) Practical
realization of the TCDA at Ka-band; 3) Size, weight, and power (SWAP) performance, and 4) The complexity of the feeding network.

To address these challenges, we first introduce TCDA phased array designs using vertically oriented dipoles that can be practically implemented using conventional printed circuit board techniques. The coupling between adjacent elements is used to significantly increase bandwidth while keeping the overall height of the array above the ground plane.

To save time in the design process, a simple circuit model is used initially to extract the design parameters of the unit cell. We improved the accuracy of the previously published circuit model by adding the model of the feed gap. Full wave simulations are used to validate the circuit model by using Ansoft HFSS software.

Furthermore, we co-designed the feeding network with the phased array antenna to achieve lower loss and higher compactness. Up to date, researchers have focused either on phased array design or on feeding network design. They have never addressed the whole system as a single design; therefore, any possible chance of collaboration or simplification between the feeding network and array antenna is diminished. In this work, we reduce the size and loss of phased array co-designing the arrays with the feeding network. Firstly, we adapted optimum beam forming architecture. This architecture utilizes phase shifter at each unit cell and true time delay at the sub array level. Secondly, we propose to integrate compact phase shifters into each array element of the TCDA using low-loss, low-profile micro-electro mechanical systems (MEMS) technology. This approach is effective in integrating the beam former into the radiating aperture without changing the overall thickness of the array. The phase shifter is designed by using circuit
simulation tools such as microwave office and ADS. Subsequently, a full wave simulation of the phase shifter is performed by using Ansoft HFSS.

The fabrication challenges that arise due to the smaller features for Ka-band operation, such as the minimum trace width, and the size of the feeding cables and connectors, are addressed by performing the beamforming at the unit cell level, while combining several unit cells through the same connector and feeding cable. The performance of the unit cell with the integrated phase shifter is examined using the full wave simulator Ansoft HFSS. Finally, minimum cost requirements are addressed through the design simplicity, in which, vias, and multi-layer structures are avoided, while the minimum requirements of the affordable PCB fabrication technique are satisfied. The proposed Ka-band TCDA array with integrated MEMS phase shifters is 4.4mm in overall height and covers 18-40GHz continuously with ±45 degrees scan capability.
Dedicated to my family
Acknowledgments

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Chapter 1 : Introduction

1.1 Phased Array Antennas for Mobile and Vehicle Mounted Applications at Ka-Band: Motivation, Challenges and Objectives

In the 1970s, it was recognized that the current frequencies used by satellite communications (SATCOM) would not satisfy the bandwidth requirements for fulfilling the anticipated demand for service [1]. There has been fast growing need for cellular and personal communications (PCS) services and increasing demand for exchanging high-resolution video and other types of data in commercial and military applications. Thus, the move to higher Ka-band for SATCOM and 5G commercial applications fulfilled this need [2].

Today’s military systems collect and process a huge amount of data from different devices and sensors that are distributed among equipment, manned and unmanned vehicles, ships, and personal, and then redistribute the information to an end-terminal. This connectivity provides the modern warfighters with significant awareness and informational superiority. Due to size, and power limitations and cost considerations, there has been interest in a common set of broadband antenna that replaces the increased number of topside antennas and RF sensors. One of the significant trends is the advanced multifunction RF systems (AMRFS) program was launched in 1996 [3] [4].
Moreover, the high levels of informational superiority and situational awareness in battlefields have led to an increased interest in electronic scanning antennas (ESA). Smaller and more agile missiles operating in non-conventional warfare has required higher scanning speed—as well as reduced scanning power consumption and footprints. Unmanned aerial vehicles (UAVs) or fast moving tactical missiles require fast scanning and rapid scene correlation to deliver early warning of attack, target detection and recognition, target tracking and countermeasure decision. These created the need for electronic scanning antennas (ESA) with rapid beam pointing feature without mechanical manipulation.

Developing the next generation of wireless systems at Ka-band that can keep pace with the high performance of SATCOM applications is not an easy task. Platforms such as wireless sensors, unmanned vehicles or hand-held devices have limited size, weight, and specific power budgetary, yet they still must be equipped with multi systems and functionality and provide high bandwidth connectivity. Additionally, the vast distribution of these functionalities across large number of devices and users required low-cost implementation. Therefore, there has been significant interest in developing wideband, electronic scanning antenna (ESA) at low profile, simple, and costly effective implementation for the next generation of SATCOM applications at Ka-band. Of particular interest are wide-band low-profile phased arrays that can steer the beam electrically and cover wide scanning field of view (±45°). Phased array is ideal for SATCOM applications since it meets service requirements and offers low-profile cost-effective solution. Phased arrays can be mounted conformally on the skin of a vehicle.
Below, we present a brief survey of existing phased array technologies for Ka-band coverage.

1.2 Survey of Existing Beam-Steering Technologies at Ka-band:

Since the evolution of mobile and wireless communications, there has been much interest in developing agile beam steering technology. With the emergence of new capabilities and requirements of SATCOM applications, there has been wider interest in a high performance beam-steering antenna. The following is a brief introduction to the development of four major beam steering techniques with their capabilities and significance to SATCOM applications.

Multi-beam array using end-fire unit element:

One of the earliest and simplest beam steering techniques is the switched beam array, shown in [5]. Beam steering in this type of array is based on switching between the elements that are tilted in different directions for a certain beam scan angle. Although this type of array is made of highly directive antenna, a dielectric lens is usually placed at the middle for increased gain. In general, this technique is simple and satisfies the required link budget. The design shown in [5] is one of many good examples, which consists of Yagi-Uda radiators and offers 22dB of gain. The entire dimensions are less than 10cm × 10cm at 24GHz. However, regardless of its attractive features, this design suffers from limited beam scan coverage. To compensate for this limitation, several units can cooperate for full scan coverage; however, this would increase significantly the size, weight and complexity of the mounted vehicle. This might be impractical for most of the SATCOM applications. Thus, a more agile beam steering technique is desired.
**Microfluidics-based beam-scanning array:**

A new technology that has recently appeared is the microfluidics-based beam scanning. It relies on a microfluidic metal that is circulated to direct the beam at a certain direction. A hemispherical dielectric lens to achieve higher gain usually backs the array. This technique is promising for low-cost realization of high-gain beam-scanning antenna array. One good example is the micro-fluidic based Ka-band beam-scanning focal plane array that is proposed in [6]. This array provides ±30° beam-scan and >21dB of realized gain at 30GHz at low profile (~10x10cm). The drawbacks of this technology are the extremely narrow bandwidth 3.3%, and relatively slow beam steering.

**Hybrid mechanical-electrical beam scanning:**

A better approach that addresses the beam scan limitations of the previous techniques is the hybrid mechanical-electrical beam scanning [7]. This design utilizes electrical switching and mechanical rotary to provide full scan coverage. It consists of array panels that are tilted vertically at different scan angles while the panels are placed on top of a rotary positioner that provides continual scan in the azimuthal direction as shown in Figure 1.1. Due to this capability, hybrid mechanical-electrical beam scan antenna has been the conventional approach for beam steering in many SATCOM applications such as movable vehicles for sensing, scanning and navigation.
Yet the design is bulky, heavy and large. The design weight and size can reach 12.5kg and 60x20cm at Ka-band [7]. This makes them impractical for SATCOM applications especially low profile applications at Ka-band. Moreover, fast scanning and rapid scene correlation for fast target detection and recognition demands rapid beam pointing without mechanical manipulation.

**Electronically steered phased arrays:**
Phased array is composed of large number of radiating elements. This type of transmission was firstly proposed by K. F. Braun in 1905 [8]. The beam in phased array is electronically steered by changing the phase of the fed current at each element. Phased arrays are popular of their high gain, directional beam, and high-resolution sensing and fast scanning capabilities. As interest in electronic scanning antennas grew, phased arrays have developed significantly, leading to many design approaches with varied features and capabilities. Such as on-chip phased array antenna [9] [10], and the tightly coupled phased array [11]. While the on-chip antenna is costly- and immature, tightly coupled
arrays have shown impressive wideband performance at extremely low profile as will be explained in more detail in chapter II.

1.3 Phased Arrays:

Phased array consists of large number of elements. The overall radiation of conventional phased arrays equals the sum of radiation pattern of each individual element (assuming no coupling). The radiation pattern can also be expressed in terms of two factors, element pattern and array factor.

The array factor, in general, is a function of the number of elements, their geometrical arrangement, their relative magnitudes, their relative phases, and their spacing, whereas, the element pattern relies more on the antenna type and its specific features. In case there is significant mutual coupling between the elements in the array environment, the element pattern of each element varies from the other elements depending on its position in the array and some other factors.

The radiated pattern of an arbitrary array at the far field can be expressed as:

\[
E(r) = \frac{\exp(-jkr)}{R} \sum a_i f_i(\theta, \phi) \exp(jkr_i, \hat{r})
\]  

(1.1)

Where, \( f_i(\theta, \phi) \) is the vector element pattern, \( a_i \) is the applied element weights, \( R \) is the far-field distance from the center of the array, and \( r_i \) is the element distance from the center of the array. The scalar array factor equals:

\[
F(\theta, \phi) = \sum a_i \exp(jkr_i, \hat{r})
\]

(1.2)

Moreover, complex weights \( a_i \) equal:
\[ a_i = |a_i|\exp(-jkr_i \cdot \hat{r}_0) \] (1.3)

Where, \( \hat{r}_0 = \hat{x}u_0 + \hat{y}v_0 + \hat{z}\cos \theta_0 \) and \( k = \frac{2\pi}{\lambda} \)

These weights are responsible for beam steering and directing the beam peak to an angular position \((\theta_0, \phi_0)\). At this angle, the exponential term in equation (1.3), cancels the exponential term in equation (1.2) and the array factor is the sum of the weight amplitudes.

In most applications, avoiding multiple maxima in the radiation pattern is a requirement. These maxima beside the main maximum are referred to as grating lobes. The grating lobes are fundamental aspects in all phased arrays regardless of the type or shape of the antenna. Eventually, grating lobes consume the energy, and increase the cross-polarization level, thus, they should be avoided. Inter element spacing is a key factor in avoiding the grating lobes. For instance, in broadside uniform array with no scan, a maximum separation of one-wave lengths is desired in order to avoid the grating lobes [12], i.e. \( d_{max} < \lambda_{high} \), where, \( \lambda_{high} \) is the wavelength at the maximum frequency of the band. Figure 1.1(a) shows the array factor at different element spacing. At one wavelength of spacing between the adjacent elements, two grating lobes appear at the end-fire direction.

In scan case, the appearance of the grating lobe is related to the scan angle. The minimum element spacing is reduced as the scan angle increases. The following relation describes the element spacing limit of uniform phased array in case of beam steering at angle \( \theta_o \) from the broadside.
According to equation (1.4), the minimum desired spacing between the elements in case of $60^\circ$ beam scan reduces to $0.54\lambda_{\text{high}}$. In other words, a uniform array with element spacing of $d_{\text{max}} = \lambda_{\text{high}}/2$ has a maximum capability of steering the beam up to $\sim60^\circ$.

Figure 1.2 (b) represents the array factor of a uniform array at different scan angles. After $60^\circ$, the grating lobe level becomes so significant. At $90^\circ$, the grating lobe has the same amplitude as the main beam level.

From the previous relations, we coinclude that small cell size is desired to avoid the appearance of the grating lobes inside the visible region of the antenna. In practice, too small unit cell size is not preferred. Since, small cell size leads to over populated aperture that is difficult to feed and excite, as well as, costly and complex to design.

Besides, small radiator size that corresonds to the cell size usually have high radiation

\[
d_{\text{max}} \leq \frac{\lambda_{\text{high}}}{1 + \sin \theta_o}
\]  

(1.4)
resistance value that is complex to be matched with a 50 or 75 ohms feeding lines. Thus, the ideal design is the one that compromise between these two factors.

In practice, phased arrays comes backed with a ground-plane, that is sometimes forced by the hosting platform, or added to ensure a single beam and to reduce the RCS of an active array. After examining the radiated field behavior over frequency, we notice that the radiated field attenuates significantly at certain frequencies. These frequencies correspond to array height above the ground plane by multiples of $\lambda/2$. The reflected field from the ground plane at these frequencies is added destructively with the original radiated field leading to poor radiation. This behavior is explained by using the waveguide model that is proposed in [13] as shown in Figure 1.3. The impedance of the lower waveguide, $Z^{1+} = jZ_o \tan \beta h$, reaches zero whenever the height ($h$) is multiple of half-wave lengths as shown in Figure 1.4 (a). This leads to a high reflection at the feeding port.

![Diagram](image)

**Figure 1.3:** (a) Phase array above ground plane, (b) Corresponding equivalent circuit model as proposed in [13].
A proof-of-concept full wave simulation of a dipole array at height $h=3\text{mm}$ above the ground is performed. As shown in Figure 1.4 (b) the array is short-circuited at 50 and 100GHz, where the height $h$ is multiple of half-wave lengths.

![Figure 1.4: (a) Draw of the impedance of the shorted waveguide ($Z^{1+}$). (b) The VSWR of the unit cell with $h=3\text{mm}$.](image)

1.3.1 Wideband Phased Arrays above a Conducting Ground Plane

The demand for wider bandwidth is increasing at an exponential rate. Web browsing, high quality video streaming and bulk data transfer are growing rapidly, and with the advent of cellular communications, smart phones and mobile broadband, traffic on mobile and wireless networks is growing at an even faster rate.

Furthermore, today’s military systems collect and process a huge amount of data from different devices and sensors that are distributed among equipment, manned and unmanned vehicles, ships, and personal, and then redistribute the information to an end-terminal. This connectivity provides the modern warfighters with significant awareness and informational superiority. However, the increased number of functionalities and
topside antennas on a limited budget, low-profile platforms, has created the interest in replacing these antennas with a common wideband set. Besides, the vast distribution of these functionalities across large number of devices and users required low-cost implementation. Therefore, there is significant interest in developing wideband, electronic scanning antennas (ESA) at low profile, simple, and costly efficient implementation. Of particular interest are wide-band low-profile phased arrays.

Backed phased arrays suffer from limited bandwidth as described in previous section. It is shown that, ground plane acts as a field reflector that deteriorates the radiated field at certain frequencies. This creates upper band limit of the array. Whereas, at lower frequencies, the band is limited by the ground plane as well. As in low frequency, the array height appears electrically small, and the array is shorted by the ground plane.

Several wideband technologies have been developed to address the needs described above using a variety of different techniques. In the following section, we present a brief introduction to a number of the most popular types of wideband phased arrays, their operational principle and their relative weaknesses and strength.

**Tapered Slot and Travelling Wave Arrays**

As in Figure 1.3 (b), the bandwidth limitations of ground-backed arrays results from the impedance of ground plane in parallel with the radiation resistance. For wider bandwidth, the reactance seen at the feed point should be minimized. This can be accomplished through minimizing the effective radiation resistance, or increasing the inductance of the ground plane. In practice, this can be performed by placing tapered transmission line between the feeding points and the top of the array to reduce the radiation resistance. The
most basic implementation of this concept is the Vivaldi array that uses an exponentially tapered transmission line that is fed by a Marchand-balun as shown in Figure 1.5 (a). Vivaldi arrays have been used for more than 40 years [14] [15], and still among the popular wideband phased arrays.

![Vivaldi Array Diagram](image)

Figure 1.5: (a) Flared-notch (Vivaldi) array [16]. (b) Banyan tree antenna (BTA) [17]. Resulted in Vivaldi arrays that achieve ultra-wide bandwidth in excess of 10:1 at wide scan up to >45° from broadside [18].

As effective bandwidth of the impedance transformation is determined by the length of the tapered transmission line, usually Vivaldi arrays are high. Their thickness is commonly 2-3λ at the highest frequency. Therefore, a major drawback of the Vivaldi array is their large size. Additionally, Vivaldi arrays suffer from high cross-polarization while scanning. This is due to the strong vertical current running down the length of the tapered line, perpendicular to the plane of the array. Leading to high cross-polarization level that might exceeds the co-polarization levels at small scan angle as 30° [19]. Moreover, Vivaldi consists of deep, vertically integrated notches, that should be
connected to avoid resonances [20], resulting in complicated dual-polarized array that is costly to manufacture and assembly. Modular Vivaldi array variations, such as the Mechantoch array [21], or the body of revolution (BOR) Vivaldi array [22] exist, but they rely on special fabrication machining and are difficult to implement at high frequencies (> X-band).

Vivaldi arrays have been used in many wideband applications, from radar to radio astronomy. However, their significant large thickness and poor polarization purity have limited their usefulness applications. Thus, recent focus on alternated low-profile wideband array technologies has remained motivated.

Several arrays have been developed as low profile and modular alternatives to Vivaldi arrays. These arrays consist of vertically integrated PCB cards backed with a ground plane. They usually <\lambda_{high}/2 depth, independent on frequency and don not required electrical connection. The balanced antipodal Vivaldi antenna (BAVA) [23], doubly-mirrored BAVA (Dm-BAVA) [24], the bunny-ear [25], and Banyan tree antenna (BTA) [26] [27] [17] arrays belong to this class (See Figure 1.5 (b). These designs represent an improvement over the standard Vivaldi design that mitigates some of the above drawbacks. Rather than using simple flared slots, the radiator in Bunny-ear array is shaped like a pair of “bunny ears” that provides a smooth transition between the stripline feed and the radiator. It has demonstrated a 5:1 bandwidth, using external balun and resonance-suppression resistor between its elements arm and ground. The Dm-BAVA, eliminated these resonances by using element mirroring in E- and H- planes, but requires 180° hybrids to maintain beam collimations. The BTA array has reached up to 4:1
bandwidth- without balun or hybrids- using shorting strips between the ground plane and the arms. This reduces its total height. Moreover, the shape of the radiator has added additional degree of freedom allowing further reduction in thickness that is typically \( \sim \lambda_{\text{high}}/2 \) tall.

However, this low profile and reduction in thickness has come at the expense of reduction in of impedance bandwidth compared to the Vivaldi. More recently, a BAVA array with 10:1 bandwidth has been demonstrated, albeit with high impedance mismatch, VSWR\( \leq 4 \), when scanning to 45\(^\circ\) [28]. Moreover, vertically integrated arrays are difficult to integrate in dual-polarized structure, especially at high frequencies, thus, it is desired to have fully planar wideband topology.

**Tightly Coupled and Connected Arrays**

While vertically integrated arrays are difficult to integrate in dual-polarization topology at high frequency, fully planar arrays have become desirable. Today’s several quasi-planar wideband arrays have been announced. Connected arrays and tightly coupled arrays make use of horizontal dimension using the mutual coupling. It might be seen unusual to benefit of the mutual coupling, however, it is clear to figure that low-profile array with no mutual coupling can never achieve wideband.

The principle behind the connected array is to achieve frequency-independent behavior through the interaction between the element and its neighbor, in a way that the element appears much larger than its electrical length and thus, supporting much larger wavelength. The idea of using connected elements to improve the low-end bandwidth was suggested as early as 1970, by Baum [29]. Later, the idea is developed by B. Munk
Munk realized that if there was a way to realize Wheeler’s current sheet, such an array would support very large bandwidth and excellent scanning performance. He started his implementation with small dipoles that are capacitively coupled at their tips as shown in Figure 1.6. He called his design the current sheet array [32] [33] [11].

Figure 1.6: (a) Tightly-coupled phased array above ground plane. (b) Munk’s current sheet array (CSA) with interdigital capacitors in dual-polarized setting [11].

Later, these types of arrays were referred to by tightly coupled array [34] [35]. When the neighbor radiating elements are connected with each other’s rather than capacitively coupled, in this case the array is called connected array [36] [37].

In general, tightly coupled phased array above ground plane is band limited by the existence of the ground plane. When the height of the array reaches multiple of $\lambda/2$, the image current will interfere destructively. Similar to all planar arrays without grating lobes, the ground plane in tightly coupled arrays can be modeled as a short-circuited
transmission line in parallel with the radiation resistance. The inter-element capacitance in tightly coupled dipole array is modeled by a series capacitance. Accordingly, the resulting load impedance is as shown in Figure 1.7.

![Equivalent Circuit of Tightly Coupled Phased Array](image)

Figure 1.7: The equivalent circuit of the tightly coupled phased array.

The inter-element capacitance along with the self-inductance of the dipole compensates for the reactance of the transmission line. In other words, the L-C might be thought of as a simple impedance matching network of the transmission line load. That is how Munk was able to achieve almost 5:1 bandwidth with his tightly coupled dipole array (TCDA), at VSWR<2 at broadside [11]. In contrast, connected dipole array are capable of only 1.5:1 bandwidth when placed above ground a plane [36].

Besides dipole, many other elements can used to create current sheet. The interwoven spiral array (ISPA) uses tightly coupled spiral elements to achieve 10:1 bandwidth at VSWR<2 without the use of any material loading. Slot arrays have also been used in tightly coupled array design [38] [39]. There is also fragmented array type that uses computer algorithm to design the shape of the radiating element [40] [41]. Such arrays have yield 8:1 bandwidth with ≥3dB mismatch loss when it is above ground plane [40].
1.3.2 Feeding Network and Beam Scanning in Phased Arrays.

There are several array architectures that describe the ways to combine the electromagnetic elements, aperture, control and power division in an array [42]. Architectural choices start from the apertures, and determine how the elements are to be clustered and fed. Behind the aperture there are ways of time delay or phase control, followed by a distribution network that gathers all the power from various elements, provide amplification, amplitude weighting, time delay and maybe adaptive control for noise suppression. The control process starts with the phase shifter component that has been the spine of electronic scanning systems since the first array was built. This component has been developed recently, to address the needs of wideband and highly flexible arrays.

Figure 1.8: Basic array construction: (a) “tile” construction, and (b) “brick” construction.
Array cost is considerable factor in many applications. There has been a need for optimum array architectures, optimum choices for collecting, fabricating and assembling array elements and corresponding array feed that is compatible with various ways of grouping elements. Figure 1.8 shows the brick and tile construction was of an array [43]. In the tile construction, or what is also called monolithic array constructions, shown in Figure 1.8 (a), the array consists of multi-layers. Usually, in this type of array, radiator patches are commonly used, as well as dipole, and fed by microstrip transmission lines or other planar feed lines. Tile construction is mainly useful when the array is to be uniformly illuminated. On the other hand, brick construction uses the vertical (depth) dimension to provide the required functionality. It can be fabricated by monolithic integrated circuit technology, thus, compatible with low-cost production. The radiator elements used in brick construction, such as dipole and bowties, have usually broader bandwidth than microstrip. Brick fabrication is generally preferred as it provides more room for power divider, phase shifter, and other components.

**Electronic Beam Scanning in Phased Arrays**

In phased arrays, beams are scanned by shifting the phase of the signal from each radiating element. Constructive/destructive interference leads to beam steering in the desired direction. According to the radiated field equation at the far zone, equation (1.2), the complex weights $a_i$ are responsible for beam steering and directing the beam peak to an angular position $(\theta_0, \varphi_0)$. At this angle, the exponential term in equation (1.3), cancels the exponential term in (1.2) and the array factor is the sum of the weight amplitudes.
This weight, gives stationary pattern peak at all frequencies. In practice, this relation is achieved by true time delay that carries linear phase relationship with frequency. However, if this condition is not met for certain reason, in this case, the weight will have the following format:

\[ a_i = |a_i|\exp(-jk_0r_i.\hat{r}_0) \]  

(1.5)

Where, \( k_0 = 2\pi/\lambda_0 \). Moreover, the new array factor equals:

\[ F(\theta, \phi) = \sum_i |a_i|\exp[jk_0(r_i.\hat{r} - r_i.\hat{r}_0)] \]  

(1.6)

This equation indicates that the array factor is a function of \( \hat{r}_0 \) and \( \hat{r} \). In other words, if the array scans at certain angle the pattern would remain unchanged except for a translation. For instance, let us consider phase scanning in one dimension. At this case, the complex weighting has the exponent:

\[ \exp\left(-\frac{2\pi}{\lambda_0} nd_x u_0\right) \]  

(1.7)

Moreover, the value of \( u \) corresponding to beam peak is given:

\[ u = u_0 \frac{f_0}{f} \]  

(1.8)

This results in pattern squint in which the beam peak angle is increased for frequencies below the designed frequency and reduced for frequencies above the design frequency. Beam squint becomes more significant as the scan angle is increased or as the array is made larger.

**Exciting the Phased Currents for Beam-Steering**

There are different ways to deliver the desired current to each element of the phased arrays. Space-fed active lens, shown in Figure 1.9 (a) and reflect arrays [44] [45] are
popular approaches to provide beam steering phased array. They rely on a single antenna that illuminates an aperture the aperture of the phased array. In this array, each element is attached with a phase shifter that point to the desired direction. This topology significantly reduces the cost and weight of the system by eliminating the need of corporate feeding. Thus, they gained their popularity in space-based radars and ground-based arrays. However, this approach suffers of instantaneous bandwidth that is limited by the use of phase control at the illuminated aperture.

Another approach is using the matrices array. This array produces orthogonal sets of beams with uniform aperture illumination. A particularly impressive implementation is the Rotman lens [46] that provide good wide scan angle that exceeds 45°. Figure 1.9 (b) presents a draw of Rotman lens, showing several ray paths.

![Figure 1.9](image)

(a) (b)

Figure 1.9: Array feed networks: (a) Space-fed lens array, and (b) The Rotman lens.

Through the lens, and the corresponding wave front. Lens and reflector arrays have been used in satellite communications systems to reduce cost and weight of the systems by eliminating the need for complex feeding network.
**Beam Steering Modalities**

The beam squint phenomenon dictates the need for using time delay steering for large arrays with modes bandwidth and wideband arrays. Therefore, it is desired to use constrained feed, such as corporate or series feeds. Compared with the series feed, the corporate feed guarantees equiphase signal distribution for wideband arrays. Furthermore, it produces beams with frequency-independent scan angles.

Mainly, there are three different technologies for controlling beam steering in phased arrays, which are optical, analog and digital. The suitable control design to be used is determined depending on the systems requirements and physical contains.

Firstly, in optical controlling network, the optical signal is modulated by an RF signal, and the optical power is divided into different channels, and then it is time delayed by fiber time delay unit before it feeds the elements. The optical technology can provide accurate time delay with negligible dispersion. Therefore, it is suitable for large arrays with wide bandwidth. However, the optical path is inefficient, as it requires amplification elsewhere in the network, also, optical control networks are usually complex. The primary obstacles to widespread use of photonic array control are network losses and device size constrains.

Secondly, digital beam forming utilizes amplification at each element in RF stage as shown in Figure 1.10 (a). Then the signal from analog to digital using A/D converter is converted. At digital domain, the amplitude and phase weights are performed; this creates an accurate pattern control and the applicability of adaptive processing. Digital beam forming gives optimum antenna performance. It can provide multiple simultaneous
beams, perform time delay and wideband operations, offers failure detection and correction for arrays and provide separate control for each path through the array. However, digital beamforming remains impractical for large arrays due to limiting factors such as, computer speed and storage requirements, A/D and D/A bandwidth, power requirements, and size.

Finally, analog controlling technology is the most mature among the three. It is a function of the well-established time-microwave analog technology and it is still advancing fast with solid-state devices integration. Figure 1.10 (b) represents simplifies block diagram of analog control array. As shown, phase shifters or time-delay devices scan the beam, while, corporate power divider network feeds the element. Different weights can be applied to achieve low side lobe level. The drawback of analog control is that it suffers of high loss due to the circulator, phase shifter, power divider and control devices. Therefore, it is common to use solid-state T/R modules at each element, or at sub array level to reduce the loss [42].

In conclusion, optical control is suitable for large wide band arrays; however, it is lossy, complex and large. Digital control is limited to small narrow band arrays. Whereas, analog control is suitable for wideband arrays and the loss can be mitigated if solid-state technology is used combined with the right design architecture.
Figure 1.10: Block diagrams showing the main differences between (a) digital beam former, and (b) analog beam former.
RF Components for Array Control

To date, phase shifter has been the most important component used. The first components that were used for phase control are the waveguide ferrite phase shifters. Phase shifters can be classified as analog phase shifter, that are controlled by voltage or pulse length, and digital, that carries small number of phase shift states. Ferrite phase shifters have been the common mean of phase control in radar and high power applications [47] [48]. They have the capability of handling hundreds of watts of power. In addition, ferrites can offer wide range of switching speeds. However, most of the ferrite phase shifters are non-reciprocal, and they require additional operations to return the phase shifter into its base state after shifting. In addition, ferrite phase shifters in general are heavy, bulky and require huge power when compared to solid-state phase shifters. This leaves ferrite phase shifters applicable for large applications but not for low profile applications and airborne.

On the other hand, solid-state phase shifters come at much lower profile. These days, there are large varieties of phase shifter circuits. Most commonly, the switched line phase shifter, that consists or multi lines with different phase delays. There are also hybrid, loaded line and reflection-line phase shifters.

Solid-state phase shifters have utilized diodes. This offers low loss and high switching speed for applications below 2GHz [49]. However, they desired high dc bias power that is unacceptable for some applications. PIN diode phase shifters was built with low loss (<1.5dB) up to Ka-band. Varactor diode phase shifters shows minimal control power consumption at the expense of higher insertion loss.
Micro electro-mechanical systems (MEMS) technology has been used in phase shifter design. They come in the same circuit topologies as described in diode phase shifters. They add lower loss, lighter weight and much lower power consumption among diode and MMIC phase shifters. Accordingly, they play significant role in large arrays. Yet, there are still development needs at the reliability limits and a need for improved insertion loss, especially at frequencies larger than Ka-band.

1.4 Limitations and Challenges in Implementation of Wideband Phased Array at Ka-Band

While wideband phased arrays have attracted the attention as the new generation of beam steering antenna at Ka-band, there are major challenges in these arrays and in their implementation at the higher Ka-band.

1. Implementation of the tightly coupled dipole array at Ka-band.

The tightly coupled dipole array with integrated balun has been implemented over 0.28-5.9GHz [50]. This design is not scalable to Ka-band and above. Due to fine features of the design, material loss and physical size of the array, realization of the design through the standard PCB fabrication techniques is inapplicable. Therefore, it is desired to overcome these limitations, and to implement the tightly coupled dipole array with integrated balun at the Ka-band at simple low-cost realization. Moreover, current design procedures of tightly coupled dipole array relies initially at circuit simulation tools that leads to the optimum design in a timely manner. This step reduces the cost and time significantly. Then the performance is verified and modified by using full wave simulators that usually take much longer time however,
gives results that are more accurate. Currently, the equivalent circuit that has been used to model the tightly coupled dipole array is the one that is proposed by Munk [11]. This circuit model was built upon the assumption that the dipoles are equivalent to short wires, therefore, the feed gap of the dipole is neglected. This circuit model can predict the behavior of the full wave simulation up to some extent. However, in beam scan scenario, the accuracy of the circuit model degrades. Therefore, the achieved result from the circuit model is unreliable. Since, the full wave modelling is costly, and takes long time, a reliable circuit mode is desired.

2. **Demand of agile wideband beam forming.**

Beam steering requires corporate feeding architecture of the phased array. Yet, there are three main beam-forming modalities, digital, optical and RF beam forming. While digital beam forming provides high-resolution beam scanning, low loss, and signal refinement for each channel, to date, it is limited to narrow band applications. Similarly, optical control beam forming is complex, lossy and bulky. RF beam controlling is mature modality. It is suitable for wideband and large phased arrays. It improves the size and loss of the control network. Yet, it suffers of relatively high loss. Furthermore, since RF beam controlling relies on phase shifters and true time delay units for beam steering, a corresponding wideband performing phase shifter is desired for wideband array performance.

3. **Discrepancies of the feeding network.**

Previous designs of the phased arrays focused on the array aperture design. None has
discussed the feeding network as part of his design. Feeding network of phased arrays suffer from many discrepancies, such as complexity, loss, weight, size and power consumption. Figure 1.11 presents an example that shows the complexity and large size that is embedded in the power division network of an $8 \times 8$ phased array. To date, phase shifter has been the most important component used. Due to the large number of phase shifters in the array, developing low-loss, low power consuming and low profile phase shifter, improves the performance of the feeding network. In addition, simplifying the connectivity and the power division network leads to a significant improvement in feeding network performance as well.

![Diagram of feeding network](image)

Figure 1.11: Feeding network of an $8 \times 8$ tightly coupled dipole array [51], showing the complexity, large size and weight of the power division network.
4. **Realization of the tightly coupled array at Ka-band.**

Durable material that withstands the corrosion, wear and tear is significant in SATCOM systems. Ceramic materials in general provide this feature and have more sustainability against the environmental reactions among other substrates such as Rogers materials, while offering extremely low material loss. Additionally, due to the small feature sizes and extremely thin layers of the array design, strain relief of the structure has become a challenge. Feeding each unit cell through a connector is impossible. On chip antenna is an approach that could overcome these limitations. However, to day, on-chip technologies are extremely expensive. Thus, compromising between the compatibility of the on-chip technology and the low realization cost of the PCB fabrication is desired.

**1.5 Contribution of the Dissertation**

The key contributions of this dissertation addressing the aforementioned challenges include:

1. **Development of the single-layer vialess tightly coupled dipole array with integrated balun at Ka-band.**

Tightly coupled dipole array at Ka-band is designed. Initially, an accurate full circuit model of the tightly coupled dipole array with integrated balun is developed. In this model, the feed gap is taken into account as inspired by [52]. Then, the array is constructed based on brick construction that allows more space for integration of control components inside the structure of the array. The dipole components are printed on low loss Rogers 3010 substrate with loss tangent ~0.0013 @10GHz.
array and its feed are designed on a single dielectric layer with double-sided metallization to reduce the complexity and cost of the design. Also, via structures are voided. To reduce the effect of leaky-wave radiation from the unshielded parts, such as Marchand-balun, symmetric structural is made. This symmetricity leads to field-distortion cancelation. Finally, the design is made within the fabrication limitations of the affordable PCB fabrications technique.

2. **Developing corporate beam forming approach and the utilization of a 2-bit switched network micro electro mechanical (MEMS) phase shifter.**

For wideband width, low size and beam-agility RF beam control modality is adopted. To further reduce the loss contributed by the feeding network, solid-sate phase shifter is used for beam forming. More specifically, micro-electro mechanical systems (MEMS) technology is used to develop the phase shifter. A 2-bit MEMS phase shifter with <.52dB loss at 40GHz and 1.8 ×1.8mm size is used. The phase shifter consists switched networks and MEMS capacitors for phase delays. The phase shifter is developed based on switched network circuit design leading its low profile.

3. **Integration of the phase shifters at the backside of the array unit cell.**

Conventionally, the beam-forming network comes after amplification in RF beam formers as sown in Figure 1.12 (a). However, this is physically impractical in array design at Ka-band and above due to the size limitations. The smallest connector type available in the market nowadays is the miniSMP connector with ~4.8mm width. Whereas, the unit cell size at high frequency such as Ka-band falls within several mm. In other words, it is impossible to feed each cell by its own cable. There are on
chip receivers that have been achieved based on CMOS with LTCC and LCP technologies [53] [54], these chips can be integrated with the array. However, on-chip technology is still immature and so costly. In this work, we propose a novel technique that addresses this challenge. In our technique, the phase shifter of each element is integrated at the backside of the unit cell itself. Whereas, several element in the same E-plane, where combined through the same power divider that is printed on the same PCB card layer and each layer is terminated by a connector that is connected to the feeding cable as shown in Figure 1.2 (b). As the phase shifter proceeds the LNA in this case, the phase shifter is wisely selected so that it is extremely low loss and noise figure would remain unaltered if not improved. In addition, the connecting transmission lines between the phase shifter and the antenna are eliminated and made short.

4. Developing assembly structure for strain relief of the design.

The PCB cares are structured in a manner to provide strain relief. This is achieved through the utilization of an additional layer. This layer adds no electrical benefits to the design. However, it is used to encapsulate the connectors. The feeding connectors are soldered from the top and bottom to the ground layers leading to a fixed PCB layers.
Figure 1.12: RF beam formers (a) the Conventional, and (b) the proposed model.
1.6 Outline of Dissertation

In the next chapter, the evolution of the tightly coupled array is introduced. The theory behind the tightly coupled array, the effect of design parameters on the array behavior and state of the art phased arrays are presented. In addition, literature review about MEMS phase shifters is provided. Chapter 3 represents the design work and methodology of the tightly coupled dipole array at Ka-band. It highlights the unit cell feeding as well. Chapter 4 presents the development of the feed design. It presents the MEMS phase shifter design and the integration at the unit cell, with a proof of concept simulated beam scan performance. Chapter 5 presents measurements of the 4×4 prototype. It highlights fabrication process and the assembly procedures. Chapter 6 summarizes the findings of this work, and present suggestions for future research in this area.
Chapter 2: Evolution of the Tightly Coupled Phased Array

Creating ultra-wide bandwidth phase array that is thin and compact is not an easy task. However, this challenge is addressed by other studies through the tightly coupled phased array. In conventional phased arrays, the mutual coupling between the neighbor elements is avoided to reduce the effect on a single element performance. This had led to a large narrow band array performance. Later on, ground plane is added to phased arrays for backside radiation reduction. The addition of this ground plane has carried more limitations to the phased arrays; until the tightly coupled principle is discovered. A tightly coupled phased array (TCPA) operates in a strongly coupled, yet controlled environment. The coupling between the neighbors elements when combined with the ground plane are found to create an ultra-wideband environment. Because of the strong near-field interactions between the elements, a TCPA can be viewed as a single antenna. Yet, tightly coupled phased arrays have established new frontiers in antenna design.

In this chapter, we describe the evolution of the tightly coupled dipole array and the phase shifter device as a beam controller in phased array. We first talk about the conventional phased arrays; more focus is given to their coupling aspect between the neighboring elements. Then we discuss wheelers current sheet that leads to the connected array, a special kind of phased arrays in terms of inter-element coupling. Later we discussed the performance of the connected array when it is backed by a ground plane,
the normal scenarios in real life applications. In addition, we show that the performance deteriorates severely after backing the array with a metallic layer. This is resolved later by the tightly coupled dipole array that is proposed by Munk [55]. The tightly coupled dipole array principle is presented and a better understanding to the array behavior is described through the simulation of different unit cell designs and parameters.

At the end, we discuss the phase shifter as the main component for beam steering in corporate feeding networks of phased arrays. We introduce different designs and types of the phase shifter. Most importantly, a comparison between modern solid-state and MEMS-based phase shifters is presented. Finally, we present the advantages of utilizing micro electro mechanical system (MEMS) technology in the design of the phase shifter.

2.1 Weakly Coupled Phased Arrays

In conventional phased arrays, the mutual coupling between elements was considered detrimental and thus avoided. Therefore, elements with low coupling were preferred. For instance, horn antennas were used because they have narrow beams and low side lobe levels (SLLs) and thus do not “see” each other. Moreover, typically, the spacing between the array elements ranges between \( \lambda/4 > d > \lambda \) depending on the application. Each element radiates almost as if it was in isolation and the total array far field is determined via superposition (array factor).

A proof of concept full wave simulations of a single dipole in free space and a weakly coupled array are performed. In both cases, dipole length remains the same, and an array with square unit cell of 3×3mm size is used. The separation between element tips is made large so there is no inter-element capacitance. The performance of the two prototypes is
compared through the input impedance at the feeding port of the dipole. As shown in Figure 2.1, the input impedance of a dipole radiator in a weakly coupled array environment is almost similar to the input impedance of a single dipole in free space. Both real and imaginary parts of the input impedances behave similarly. This indicates that the dipole is not affected by its being in the array environment. In addition, a ground plane is added to the array at a height equals 3mm that corresponds to $\lambda/4$ at 25GHz. After adding the ground plane, the real part of the input impedance of the array has changed as shown in Figure 2.1. At around 25GHz, where the array is $\lambda/4$ above ground, the real part shows maximum value. This is due to the constructive reflection of the ground plane at this frequency that leads to an increase in the radiation resistance value.

Figure 2.1: Input impedance at the feeding port of a dipole in free space, a weakly coupled array, and a weakly coupled array above a ground plane.
We also compared the operational bandwidth of a single dipole and a weakly coupled dipole array as shown in Figure 2.2. It is widely known that the dipole provides narrow bandwidth ~30%. It is also expected to have similar behavior for the weakly coupled array. This is confirmed by Figure 2.2 that shows approximate impedance bandwidth of ~20% of the weakly coupled array. The slight reduction is due to small mutual coupling that cannot be ignored in reality.

The bandwidth of weakly coupled arrays is also limited at high frequencies by the existence of grating lobes. That mainly happens when the element spacing is half-wavelength for end-fire scan, or one wavelength for broadside scan. Therefore, to avoid the grating lobes, we should bring the elements closer from each other. This contradicts with the “weakly coupled” requirements. Thus, the optimum element spacing was

![Figure 2.2: Voltage standing wave ratio of a dipole element in; free space, and a weakly coupled array environment.](image)
determined based on the bandwidth requirements of the required application. Later on, several narrowband electromagnetic bandgap (EBG) structures have been developed to reduce the mutual coupling [56]. Interleaving EBG structures with the array elements is shown to mitigate mutual coupling.

2.2 Connected Phased Arrays

Further researches were conducted on the mutual coupling between elements in order to improve the array performance and to increase the bandwidth. This had led to different ideas and to the evolution of the current sheet principle. Wheeler among the researcher, has realized that current on the aperture of an array can behave similar to a current sheet. A uniform current sheet is frequency independent structure. Therefore, he thought that one could achieve a wideband array by creating a uniform current on the aperture of the phased array. In the following section, Wheeler’s current sheet is described in more details. Furthermore, we present the performance of the connected array that he had developed based on his principle.

2.2.1 Wheeler’s Current Sheet Principle

In 1948, Wheeler analyzed and proposed a uniform current sheet [13] [31]. His assumption was that an infinite planar sheet of uniform current distribution radiates identically at all frequencies. Wheeler visualized the current sheet as a planar, linear array of closely spaced and small elements. He suggested that a key understanding of wave radiation from an infinite array is to consider the collected radiation from all elements. These will be discussed starting from Figure 2.3 (a) that illustrates an infinite planar dipole array. The array is assumed to be backed by an open circuit boundary, that
is half spaced, and element separation less than \( \lambda \). The array is in the operation of a plane wave normal to the array aperture. By looking at the current formed on the dipoles with

Respect to the image theory, hypothetical walls can be assumed around each dipole as shown in Figure 2.3 (a). Perfect electric conductor (PEC) walls perpendicular to the dipole and perfect magnetic conductor (PMC) walls parallel to the dipole.

Thus, each dipole can be viewed as sitting in a PEC/PMC waveguide as in Figure 2.3 (b) that supports a TEM mode (i.e., plane wave). If this waveguide is filled with air, then its characteristic impedance equals:

\[
Z = Z_0 \frac{b}{a}
\]

(2.1)

Where \( Z_0 = 120\pi \ \Omega \) is the free space impedance [31]. For a dipole of effective height \( h \) located in the waveguide, as depicted in Figure 2.3 (c), and taking into considerations the amount of coupling between the dipole and the waveguide that is modeled as a transformer in [31], the radiation resistance becomes
From Equation (2.2) several implications can be extracted. First, the highest coupling between the dipole and the waveguide occurs when the effective height of the dipole equals the waveguide height. In this case, the radiation resistance reduces to that in Equation (2.1). Second, by switching from square to rectangle unit cell, and vary the ratio between the rectangle dimensions, one can scale the ratio \( b/a \) to obtain different radiation resistances. Finally, the unit cell impedance can be used to alter the radiation resistance as well, for instance, filling the cell with dielectric reduces the resistance among the air.

Based on these observations, Wheeler considered an infinite phased array as a planar current sheet [13]. He considered densely spaced elements to operate together at much lower frequency than their isolate resonance frequencies. In this scenario, the current almost constant over the array elements, and thus, the effective length of the dipole appears to equal the height of hypothetical waveguide and one can use Equation (2.1) to calculate their radiation resistance. For the specific case of square grid (i.e., \( b = a \)), the radiation resistance becomes \( R = Z_0 = 120\pi \ \Omega \). In [13], Wheeler has also expanded the results of [31] to include scanning. Where, in the case of scanning, the hypothetical TEM waveguide become oblique to the array plane. This changes the width or height of waveguides by a factor of \( \cos \theta \), where \( \theta \) is the scan angle as shown in Figure 2.4. Accordingly, the radiation resistance act proportionally to \( \cos \theta \) for the \( E \)-plane and \( 1/\cos \theta \) for the \( H \)-plane.

\[
R^E = Z_0 \frac{b}{a} \left(\frac{h}{b}\right)^2 \cos \theta \tag{2.3}
\]
\[ R^H = Z_0 \frac{b}{a} \left( \frac{h}{b} \right)^2 \frac{1}{\cos \theta} \] (2.4)

Since Wheeler in [31] assumed half space environment of the array, single wave-guide is hypothesized. However, what exists, as a current sheet in reality is an array of dipoles in

**TE mode (H-Plane scan)**

![Diagram of TE mode (H-Plane scan)](image)

**TM mode (E-Plane scan)**

![Diagram of TM mode (E-Plane scan)](image)

Figure 2.4: Oblique flat Array and waveguide. (a) H-plane scan, and (b) E-plane scan.
free space that are connected at their tips, as shown in Figure 2.5 (b). In this case, there are two hypothetical waveguides: one below and one above the array. Each one of these waveguides is considered as a transmission line with characteristic impedance $Z = Z_0(b/a)$. Since the dipoles radiate bi-directionally into both of these waveguides, the radiation becomes $R = Z/2 = Z_0(b/a)/2$. In case of a square, the radiation resistance is $R = Z_0/2 = 60\pi \Omega$.

After the current sheet principle was evolved, it took 20 later when it is firstly implemented. The simplest way to accomplish Wheeler’s current sheet is through a large array of connected dipoles [37] that was firstly performed by Baum who is credited by Hansen. Baum recognizes that planar dipole array in free space exhibit unlimited low frequency performance.

An array of connected dipoles is designed as shown in Figure 2.6. As suggested by the current sheet model, the resistive impedance equals $60\pi = 188\Omega$ when the current is almost uniform at the dipoles which happens at very low frequency. However, the

![Figure 2.5: (a) Wheeler’s current sheet in free space. (b) Practical implementation of Wheeler’s current sheet using connected dipoles. (c) Equivalent circuit model for the connected dipoles in free space.](image)
impedance of dipoles varies quickly as frequency increases, and these dipoles achieve self-resonance when their lengths are \( \lambda/2 \). To resolve this issue, self-complementary elements such as bowtie, can be used instead of a dipole. In this case, a pure resistive impedance of 188\( \Omega \) is achieved as shown in the second and third cases of Figure 2.6 and the bandwidth is theoretically infinite as shown in Figure 2.7. In practice, the dipole array has to be infinite and residing in free space in order to fulfill the current sheet requirements and to support an infinite bandwidth.

Figure 2.6: Free Space Impedance of an infinite array of connected dipoles. Transition from linear dipole to flared dipole with ±90° to achieve a purely resistive impedance of 180\( \Omega \).
2.2.2 Bandwidth Degradation Above a Ground Plane

Wheeler’s current sheet as described in the previous section resides in free space. However, in real world applications, a metallic layer backs the phased array. This metallic layer is forced by the hosting platform or necessarily added to shield the feeding network and concentrate the radiated field to only half space. The addition of this ground plane carries more limitations to the phased array. Figure 2.8 (a) depicts the array in previous section when it is over a ground plane at height $h=3\text{mm}$. Its corresponding impedance is compared with the freestanding case (without ground plane) in Figure 2.9. As noted, the addition of this ground plane has significant impact on the array impedance. Specifically, it shorts the array at low frequency, thus limiting the bandwidth; it adds significant inductive value to the reactance at low frequencies; and it adds inductive part
at higher frequencies. This can be simply explained by referring to the equivalent circuit model in Figure 2.8 (b). In the presence of a ground plane, the bottom waveguide, that has a characteristic impedance of \( Z_0 = 2R_{d0} \), simplifies a short-ended transmission line. With an impedance of \( Z^+ = jZ_0 \tan(2\pi h/\lambda) \). Similarly, the upper waveguide is terminated by the air and has an impedance of \( Z^- = Z_0 \). Therefore, the total impedance observed at the feed becomes \( Z^+ / Z^- + jX_A \), where \( jX_A \) represents the self-impedance of the dipole.

As can be seen, the addition of the ground plane has changed the impedance \( Z^+ \) from pure resistance to reactive impedance. Thus, the parallel combination of \( Z^+ \) and \( Z^- \) has become reactive. At specific frequencies such as when the array thickness is \( \lambda/4 \) (or its multiple), the ground plane is seen as “open-circuited”. At higher frequencies, the effect of the ground plane turns into capacitive until the thickness reaches \( \lambda/2 \), it shorts out the array. Whereas, at low frequencies where array thickness is less than \( \lambda/4 \), the

Figure 2.8: (a) Connected dipole array over a ground plane. (b) Equivalent circuit model proposed in [55].
Figure 2.9 The effect of adding a ground plane layer at $h=3$ mm, to the CSA that is shown in Figure 2.6. (a) VSWR, (b) input impedance of linear dipole, and (c) input impedance of Bowtie.

Effect of the ground plane becomes inductive. Therefore, parallel combination of $Z^{-1}$ with $Z^{-1}$ results in decreased resistance and increased inductive reactance, as shown in Figure 2.9 (b) and (c), until the array is shorted out at DC ($\omega = 0$). In conclusion, the bandwidth for most practical arrays is limited by the ground plane effect at low frequencies. For connected dipoles array, the bandwidth is limited to 2.5:1, as shown in Figure 2.9 (a).
There have been many trials to mitigate the effect of the ground at the lower frequency band. One of the approaches addressed the issues by placing a ferrite layer between the array and ground plane [57]. However, this approach increases the insertion loss and the weight of the antenna.

2.2.3 Material Loading for Wideband Performance

As shown earlier, a ground plane creates severe match distribution to the array. When it is \( \lambda/2 \) above ground, it short circuit the array. One way to extrapolate this behavior, is that the reflected wave from the ground, at height \( \lambda/2 \), adds destructively with the direct radiation from the aperture, resulting in zero resultant and no radiation. Emerging from this prospective, the destructive phenomenon can be alleviated by cancelling the reflected wave before interferes with the original radiated wave. Practically, this can be done by using a resistive sheet. This resistive sheet or resistive frequency selective surface (FSS) is also called circuit analog absorber [58]. The resistive sheet can be simply a metal ring. It attenuates the reflected wave from the ground, and suppresses the destructive ground interference especially at mid of the band leading to double the frequency bandwidth. In addition, FSS geometry is polarization-insensitive, and it can be utilized in dual-polarized arrays as well. However, FSS creates huge loss in power. Almost half of the power, back radiation, is lost. Therefore, different approach is desired especially for applications with high-required efficiency. Similarly, magnetic material can be used to fill the area between the ground and the aperture as shown in Figure 2.10 (a). More detailed analysis about utilizing magnetic substrate material will be examined in section 2.3.3.
To alleviate the loss of the substrate, one can use a superstrate dielectric material as shown in Figure 2.10 (b). The superstrate basically pulls more energy upward towards it from the substrate, leading to more current being radiated instead of being lost in the substrate. The following relation explains it. Starting from the radiation efficiency,

\[ e_r = \frac{P_{\text{radiated}}}{P_{\text{accepted}}} = \frac{P_{\text{Sup}}}{P_{\text{sub}} + P_{\text{sup}}} \]  \hspace{1cm} (2.5)

Where \( Z_{\text{sup}} \) and \( Z_{\text{sub}} \) are the complex powers transferred to the complex impedance of the superstrate and substrate respectively. After substituting the power relation of the superstrate and the substrate,

\[ P = \frac{1}{2} |I|^2 R = \]  \hspace{1cm} (2.6)

And the efficiency,

\[ \xi = \frac{R}{|Z|^2} \]  \hspace{1cm} (2.7)

The radiation efficiency can be simplified to,

\[ e_r = \frac{1}{\xi_{\text{sub}} + 1} \]  \hspace{1cm} (2.8)

As noted, \( \xi_{\text{sub}} \) and \( \xi_{\text{sup}} \) depend solely on substrate and superstrate parameters. From the equation, we can see that after adding the substrate loading, that is resistive, the efficiency \( e_r \) reduces. However, the addition of the superstrate increases the \( \xi_{\text{sup}} \) and compensates for the rise in \( \xi_{\text{sub}} \). In fact, the superstrate improves the efficiency by drawing power away from the resistive FSS or substrate to the radiation load. More analysis will be presented in section 2.3.3.
2.3 Tightly Coupled Array for Wideband Beam Steering

About 50 years after Wheeler, a practical remedy for the effect of the ground plane was proposed by B. Munk [59], [55], that suggested adding capacitance between the elements of the array as in Figure 2.11(a). This capacitance has added an additional term to the impedance of the ground plane compared to the conventional case where $jX_A = j\omega L_d$. That is for capacitively couple dipoles $jX_A = j \omega L_d + 1/(j\omega C_d)$, where $C_d$ represents coupling capacitance between dipoles as shown in (b). According to Munk, this extra capacitive term is significant at low frequency and counteracts the inductance due to the ground plane, as shown in Figure 2.12 (a). Consequently, the array impedance becomes almost resistive around the low frequency, leading to a much wider bandwidth [see Figure 2.12 (b)]. In dipole arrays, the inter-element capacitance can be achieved through different forms, such as the interdigital capacitors, or by overlapping dipole arms or their
proximities. By applying this feature, Munk exposed that such dipole arrays can achieve 4.5:1 bandwidth with $\lambda/10$ thickness (at the lowest operational frequency) at very close distance from the ground plane [59], [55]. Since Munk’s dipole array extended from Wheeler’s electric current sheet, it is referred to as current sheet array (CSA). Due to the coupling between its elements, and to distinguish it from the “connected arrays” that is also originated from Wheeler’s current sheet, this array is later called “tightly coupled array” (TCA).

Figure 2.13 (a), shows three different tightly coupled dipole array designs. In all three, the coupling capacitance is kept unchanged, whereas, dipole flare angle is changed from $0^\circ$ to $90^\circ$ with $45^\circ$ step size. By looking at the corresponding input impedance of each design shown in (a), it is noted that moving from linear dipole to flared dipole leads to changes in impedance behavior. More specifically, increasing the flared angle from $0^\circ$ (Linear dipole) to $90^\circ$ (bowtie) increases the capacitance at high frequency and forces more stable impedance over wider range of the band as shown in Figure 2.13 (b).
Figure 2.12: (a) Impedance of dipole arrays at the lower frequency band. (b) Corresponding VSWR values of the arrays.
Figure 2.13: (a) Impedance of dipole arrays at three different element shapes; Linear, Flared with 45°, and flared with 90°. (b) Corresponding VSWR values of the arrays.
2.3.1 Analysis Methods for Tightly Coupled Arrays

One of the most useful concepts in electromagnetic is to calculate the radiation resistance of an antenna of a certain length and configurations. It is not an easy task, for instance for a dipole in free space, it requires solving the spherical electromagnetic wave with all complexities attached to it. The scenario gets more complex when more than one element are involved such as in directive arrays. Self-impedance and mutual impedances are desired to find the radiation resistance of each element with respect to its current. This procedure involves unknowns that are proportional to the square of the number of antennas. Different approaches have sought to simplify this relation, such as, neglecting the relation between the far elements. Other suggests that infinite array solution yields extremely simple solution for the radiation resistance of an antenna element in large array.

Infinite array solution using the waveguide model

The simplicity of an infinite array aperture comes from its radiated plane wave, the simplest case of a wave in space. The infinite array solution works well for large arrays, except for small number of elements at edges. Other than this, the infinite array is a fair approximation to the large arrays.

The infinite array solution is approached by comparison with a hypothetical case of the ordinary transmission line that transmits plane wave as described earlier, see Figure 2.3 (a). The performance of an antenna inside the arrays is examined by looking at the behavior of the same antenna while it is radiating inside the waveguide. Therefore, the entire array is compared with many such antennas radiating into contagious channels.
The field within this periodic structure is decomposed into a set of orthogonal Eigen modes that are known as Floquet modes. Any interaction between these modes and the surrounding structure can be treated by considering the coupling to these modes. If the inter-element spacing of the array is less than the grating lobe spacing, then only two fundamental modes, TEM or TE\(_{00}\) and TM\(_{00}\), propagate in free space. The remaining higher-order Floquet modes evanesce. The fields at the walls of each hypothetical waveguide are matched except within a phase difference at each corresponding point. This is useful for simulating the array periodic structure. Recent simulation tools such as Ansoft HFSS, based on the finite element method (FEM) [60] [61] [62] [63] [64] [65], provide the tool to simulate the infinite array through the drawing of a single unit cell. Thereafter, the phase difference is determined between the walls of the

Figure 2.14: Infinite array solution through the simulation of the unit cell. (a) Infinite tightly coupled dipole array over ground plane. (b) Unit cell with periodic boundary made by Ansoft HFSS.
hypothetical waveguide that are denoted by master and slave boundary conditions as shown in Figure 2.14. Since the array is assumed infinite, the field repeats itself in a systematic pattern that leads to field calculation of the entire array.

The waveguide model can be modified to calculate the radiation resistance of an array at beam scanning as well, see Figure 2.4. In this case, the polarity of the obliquely incident wave is alternated by the conductive walls of the image dipoles in each row so there is no radiation with a wave front parallel to the array (TEM mode). The possible modes of radiation desire a separation difference between the adjacent dipoles equal to and odd-integral multiple of half-wave length for constructive combination. Therefore, only the dominant (TE\textsubscript{10}) mode is considered. Each wave front is vertical and forms with the array an oblique horizontal angle. In this situation, each mode of radiation causes a pair of wave fronts. The pair of wave fronts and effective area are the two factors that change the radiation resistance in the oblique incidence of (TE\textsubscript{10}) compared with broadside incidence (TEM). The radiation resistance at oblique incidence is determined by equations (2.3) and (2.4).

Furthermore, the radiation resistance from the waveguide model can be used to determine the directive gain of the array. The power ratio of an array to a power of a single dipole approximately equal to twice the number of elements in the array, this corresponds to the total number of dipoles and their images [31]. Very large array that consists (n) number of similar elements, and backed by a ground plane have a total radiation resistance, based on equation (2.2),
\[ R_n = \frac{nR_\varepsilon h^2}{ab} \]  \hspace{1cm} (2.9)

The radiation resistance of an isotropic antenna of effective length \( h \) is,

\[ R_l = 30\left(\frac{h}{\ell}\right)^2 \]  \hspace{1cm} (2.10)

The total effective length of the antenna in the array equals \((2nh)\) that is the length of one antenna multiplied by the number of antennas and their images. Array unit current is defined as the current that develops certain field intensity in the main pattern at certain distance that is sufficiently large that any two antennas in the array have less than one radian length path difference [31]. Similar radiated power would be generated by isotropic antenna carrying current equals \((2n)\) units. This simplifies the apparent power. The directive gain of the array equals power ratio between the actual power and the apparent power, and equals,

\[ p = \frac{A}{\pi l^2} \]  \hspace{1cm} (2.11)

Where \( A \) is the total area of the array, and \( \pi l^2 \) is the area of the circle, whose radius is one radian length. In other words, the directive gain equals the area of the array over the area of the radian circle. A special case of interest is the half-wave length dipole. The gain of each dipole equals \(1.64\), whereas for the array, \(1.91n\). Similarly, for small dipole, the gain of each element is \(1.5\), and for the array is \(2.1n\).

**Finite by infinite array solution.**

An intermediate solution that compromise between the speed and accuracy of the calculated performance of a planar array can be achieved by assuming infinite extension
of the array at one dimension, whereas, using finite number element at the other dimension. Using this approach, the radiation pattern of the array can be calculated accurate in the direction of termination. In this dimension, the field is determined based on the field contribution by each element, taking into account the field at the edges, edge effect and the surface currents due to the termination. On the other hand, the array is assumed infinite at the other dimension, therefore, there is no field effect due to the edge currents. Figure 2.15 shows 8×infinite array that is simulated by using Ansoft HFSS simulator with the corresponding boundary condition on it. As shown in (a) for beam scan in \( E \)-plane, the array is terminated in \( E \)-plane, and an absorbing layer (perfectly matched layer) [66] [67] [68] is employed to mimic the radiation condition. Whereas, in (b) beam scan is examined in \( H \)-plane.

Figure 2.15: Finite “8” × Infinite array solution showing simulation setups in two scan planes; (a) \( E \)-plane and (b) \( H \)-plane.
Finite array solution using domain decomposition method (DDM)

The number of phased array elements in some applications can range from 10 to 100 element. As a result, these arrays inherent very complex and large structure making them difficult to analyze with traditional simulators such as method of moment or finite element [69]. The traditional approach for solving large phased array has been the infinite solution. As mentioned earlier, this solution neglects the field changes due to the antenna position in the array. In fact, the environment, fields, and coupling experience by each element in the array varies by its location. Lacking this information about each element introduces limitations in finite array designs.

To solve large geometries with reliable finite element approach, domain decomposition method (DDM) is established to allow the usage of a distributed network of compute nodes and forces larger block of distributed memory. It decomposes the mesh representation of a structure into a series of non-overlapping mesh domains, where each matrix is individually solved with a traditional direct matrix, and then can collectively be used as a preconditioner for an interactive matrix solution to the full model. In this solution, each cell of the array has a unique solution that is determined based on its location in the array, furthermore, the solution takes into account the effects along the edge of the array. This approach is efficient as it is not desired to solve each unit cell in parallel at once, also, further improvements are met by leveraging the repeating nature of found in the solution matrices of specific cells residing in identical environments [70]. A comparison between DDM technique and traditional solving technique is proposed in [70]. A dual-polarized Vivaldi array of 256-elements is simulated. The array required
211GB RAM and over 122 hours of simulation time on a 256GB RAM computer. Whereas, using DDM the simulation took only 69GB RAM and 12.5 hours computational time on a cluster of 24 different machines. Resulting in a simulation that is 9.8 times faster and desires 67% less RAM [70].

The DDM feature is supported in recent full wave simulators such as Ansoft HFSS v.15. Figure 2.16 shows 14 ×8 array that is simulated in Ansoft HFSS using DDM. All needed is the unit cell design. The software replicates the unit cell in each direction as specified by the user. The element excitation can be assigned in the “array properties” panel. As shown in (a) 6 columns, 3 at each side of the row, are assigned as passive elements. These elements do not radiate and they just act as dummy elements that usually used to

![Diagram of Active and Passive Elements](image)

(a)

(b)

Figure 2.16: Domain decomposition methods in Ansoft HFSS v.15. (a) “array properties“ Panel showing active and passive elements in the 14 ×8 array. (b) Full structure of the array using the DDM.
absorb the surface waves in terminated arrays. The full structure is shown with the highlighted main unit cell and replicated cells in each dimension.

2.3.2 Equivalent Circuit Model for the Tightly Coupled Dipole Array

In this section, a developed circuit model of the tightly coupled dipole array is presented. This model is significant in that it reduces the time and cost of the design process of the array. It is also significant in better understanding the operation of the array. According to [11] an infinite X finite frequency selective surface, that consists of columns of closely spaced straight wires shown in Figure 2.17 (a), can be modeled by the equivalent circuit shown in Figure 2.17 (b), in which the inductor is associated with the short wire and the capacitor is associated with the small overlapping gap between the elements. Under some conditions, an array of short dipoles that are tightly coupled to each others can be treated similar to this FSS, and the dipole elements are modeled by an inductor in series with the inter-element capacitance. These condition are specified in [11] as follows; a) the array operates either in E- or H- plane, b) the array has no grating lobes and c) it consists of short dipoles. In other words, the array should exhibit negligible cross-polarization and low surface waves. Besides, the dipoles should be short in terms of the wavelength; otherwise, the lumped model can no longer be valid.

To this point, the waveguide model that is discussed earlier, when combined with this circuit model of the aperture, results in the final equivalent circuit of the unit cell of a tightly coupled array on a top of a ground plane [see Figure 2.18]. In this circuit model, the inter-element capacitance is represented by $C_{coupling}$ and the dipole inductance is denoted by $L_{dipole}$. The superstrate, substrate and free space layers are represented by
transmission line sections with properties determined by the propagating Floquet mode within the layer. In case of broadside (no scan), the impedance can be determined by using equation (2.12).

\[
R = Z \left( \frac{h}{b} \right)^2 = Z_0 \frac{\mu_r b}{\epsilon_r a} \left( \frac{h}{b} \right)^2
\]  

(2.12)

Figure 2.17: Modeling of FSS in (a) by the equivalent lumped circuit shown in (b).

Figure 2.18 Equivalent circuit of tightly coupled dipole array as proposed in [11].
To validate the accuracy of the equivalent circuit, a TCDA unit cell is considered as shown in Figure 2.19 (a), with flared dipoles that are printed on Rogers/Duroid5880 board of thickness 254μm and dielectric constant $\varepsilon_r=2.2$. The dipole inductance is controlled by the dipole dimensions, more specifically its width, and the inter-element capacitance is controlled by the overlapping tips at the end of the dipoles. The spacing between the dipole and the ground plane ($h$) equals 2.9mm. The unit cell is square with 3mm width. Accordingly, the values of the parameter in the equivalent circuit model shown in Figure 2.18 are assigned such as; the waveguide impedance equals $377\Omega$; substrate height equals 2.9mm; while, the capacitance and inductance values were found by optimization process. The best match between the circuit and the full wave simulation, is shown in Figure 2.19 (b). By looking at the result in Figure 2.19 (b), we noticed that there is a difference between input impedance of the circuit model and the simulation. In specific, the equivalent circuit shows more inductive impedance than desired at the upper half of the frequency band. In addition, the real part of the correct impedance, which comes from the simulation, is lower than the one that is given by the equivalent circuit, especially at the middle of the band. Thus, a better modeling of the TCDA is desired.

Back to the extraction of the circuit model in Figure 2.18, the aperture of the tightly coupled dipole array is assumed equivalent to a FSS of short wires. In fact, this assumption does not take into account the feed gap. This feed gap becomes more significant as its get electrically larger, or as frequency increases. Therefore, it should be included in the circuit model for better accuracy.
Figure 2.19: (a) Unit cell of tightly coupled dipole array at Ka-band. (b) Corresponding comparison of the input impedance between the circuit model in Figure 2.18 with $L_{dipole}=1.4\text{nH}$, and $C_{dipole}=0.043\text{pF}$ and the full wave simulation.

**Improved equivalent circuit model**

In [52] the feed gap is modeled by an inductance, relates the electromotive force and changed current at the feed, in parallel with the capacitance that is related proportionally to the gap separation. This model is added to the circuit model in Figure 2.18 to reach to the improved equivalent circuit model in Figure 2.20 (a). The new model is validated and the input impedance of the new circuit model is compared with the full wave simulation. The result in Figure 2.20 (b) shows that a better match is achieved after using the new model.
2.3.3 Modelling the Materials in Tightly Coupled Arrays

An array of tightly coupled elements above a ground plane can be designed to achieve 4:1 BW at VSWR<2 [55]. In order to achieve a wider bandwidth, several techniques can be performed.

Utilizing resistive-loaded substrate

The ground plane creates short circuit when the array is $\lambda/2$ above the ground. This simplifies an upper band limit of the array as shown in Figure 2.21. This short circuit separates wide well behaved spectral region, clearly, if this limit could somehow be eliminated, the array bandwidth can be doubled and extended to well beyond a decade. One approach is to make a ground plane that would move electrically as a function of frequency with the dipole array. However, mechanically moving the ground plane is not practical for many applications. Other approach is to use a resistive loading between the array and ground plane as shown in Figure 2.22 (a). This resistive sheet attenuates the ground plane reflection response at mid-band, and hence, suppresses the destructive ground plane interferences leading to bandwidth improvement as shown in Figure 2.22 (b). The drawback of this method is that it reduces the radiation efficiency significantly, since, almost half of the power dissipates in the resistive load at the entire band.
Figure 2.20: (a) Improved equivalent circuit model; $L_{\text{dipole}}=0.52\text{nH}$, $C_{\text{dipole}}=0.065\text{pF}$, $L_{\text{feed}}=0.607\text{nH}$, and $C_{\text{feed}}=0.0106\text{pF}$. (b) Input Impedance comparison between the full wave simulation and the improved circuit model.
Figure 2.21: VSWR showing the upper limit due to ground plane short effect at $\lambda/2$ height.

Figure 2.22: (a) Equivalent circuit of a resistive-loaded substrate. (b) The effect of using a resistive sheet at $h_1=1.7\,mm$ and $R_{FSS}=300\,\Omega$. 
Use of bulk material- substrates and superstrates

The use of bulk material significantly affects the performance and behavior of the array. Two volumes can be utilized in this process. The volume between the ground plane and array aperture can be filled with substrate, and the volume above the aperture of the array can be filled with superstrate as shown in Figure 2.23. By using substrate loading, one might think of the effective electrical thickness of the array that increases and leads to better match at specific frequencies. In fact, using the substrate will reduce $Z_{\text{sub}}$ and increases the reactance part of $Z_{\text{TCDA}}$ leading to a reduced bandwidth.

On the other hand, magnetic substrate materials have an opposite effect. They increase $Z_{\text{sub}}$, which reduces the net inductance of $Z_{\text{TCDA}}$. As a result, ferrites and other magnetic material offer a very effective manner of mitigating the ground plane effects and significantly improving the bandwidth [71] [72]. However, magnetic materials are often

![Figure 2.23: Equivalent circuit of TCDA with material loading.](image)
lossy and heavy, making them impractical for many applications. Figure 2.24 presents the VSWR of the unit cell in Figure 2.19 (a) at three different substrate material loadings. The first material (in red) is a dielectric with \( \varepsilon_r=4 \) and \( \mu_r=1 \). As shown, this material has narrowed the bandwidth. Whereas, the magnetic material (black curve) with \( \varepsilon_r=1 \) and \( \mu_r=4 \) has improved the bandwidth.

In addition to substrate loading, a superstrate can be used. If a superstrate of \( \lambda/4 \) of thickness is used, it acts as a quarter-wave impedance transformer. This lowers the effective radiation resistance, making the input impedance more real and improving the bandwidth. Another way to look at the superstrate is as asymmetrically loading the top side of the array that leads more power to be radiated up and away from the ground plane. Furthermore, superstrate is found to improve the scan volume of the array through improving the stability of the impedance over scan [73] [74] [75]. Figure 2.25 presents

Figure 2.25: VSWR at different substrate loading materials.
the VSWR variations of the unit cell when the superstrate thickness is changed. Here the superstrate has $\varepsilon_r=4$ and $\mu_r=1$. As noticed, at 2.5mm of thickness, the superstrate behaves as a quarter-wave transformer and gives best impedance bandwidth.

2.4 RF Beam Steering for Wideband Phased Arrays

RF control modality offer solutions for wideband large arrays. Among optical control, it gives more compact profile, less complexity and loss. Similarly, compare to digital control, RF control provides wider bandwidth. Yet, RF controlling modality suffers of relatively high loss when compare to digital. This loss comes from the power divider network, circulators, phase delay elements and couplers. Currently, there are several RF control architectures that offer different design aspects. Depending on the application, an appropriate architecture can be adopted.
2.4.1 Time delay Units Vs. Phase Shifters Vs. T/R Module

In the first architecture, shown in Figure 2.26 (a), T/R module and a time delay unit (TDU) is used per each element. It provides very accurate time delay and bandwidth that is subject to the antenna elements design that can reach up to 10:1 in some cases. This architecture requires amplification at each element, to compensate for the loss of the TDU. For very large wide band arrays, this technique requires time delay, by the last unit, that exceeds hundreds of wavelength, resulting in large, lossy and heavy TDU. In addition, there is little room behind each element to include the TDU, and amplification. Therefore, this architecture is practical mostly for relativity small, very wideband array.

A more practical architecture is shown in Figure 2.26 (b). In this architecture, a small increment of time delay is provided at each element, perhaps up to three wavelengths. Then, after grouping several element into subarray, and perform amplification, it provides longer delays at different levels of sub arraying. In this case, wide scan can be provided for very long wideband arrays.

For large arrays with modest instantaneous bandwidth, a more practical architecture than the one shown in Figure 2.26 (b) can be followed. The obvious solution is shown in Figure 2.26 (c). It consists of using phase shifter at the element level. Then, divide each several element into sub arrays and utilizes TDU at the sub array level. This architecture is simple, easy to implement, and provides room for including TDU at subarray level. However, it can produce significant quantization sidelobes.

Finally, scanning about time delay beam position architecture provide exact time delay at only small number, two to four, of beam positions. It incorporates complete set of TDU
and complete set of phase shifters as shown in Figure 2.26(d). The scan sector is centered on specific number of true-time delay positions, then phase shifters are used to scan the beam from the time-delayed position halfway to the next time-delayed position. This architecture is suitable for wideband arrays. However, it is costly, since it requires different set of switch lines for every element of the array.

Figure 2.26: RF beam Steering Architectures; (a) T/R module for wideband small arrays, (b) Cascaded TDU for wideband large arrays, (c) Phase shifters for large arrays with modest bandwidth and (d) set of TDUs and set of phase shifters for wideband arrays with fixed small number of beam position.
2.4.2 Micro-Electro Mechanical System (MEMS) Technology in the Design of the Phase Shifter

The beam squint phenomenon dictates the use of true time delay in wideband phased arrays. True time delay conveys linear phase delay relation versus frequency. Phase control elements are currently based on different topologies; ferrite materials, FET switched or BIN diodes. Solid-state phase shifters are compatible in shape with microwave components. They consume less DC power than ferrite and provide lower loss. In specific, FET-based phased shifters can be integrated with amplifiers on the same chip, therefore, reducing the assembly cost of the phased array system [76]. As they offer smart planar solution at microwave and millimeter-wave frequencies and they have been used largely in modern phased array systems.

Analog solid-state phase shifters are achieved using Varactor diode that provides continuously variable phase shift. Digital phase shifters give discrete set of phase delays and are implemented by using switches. PIN diode consume more DC power than FET-based ones, however, it provide lower loss especially at high frequency. Modern solid-state phase shifters are quiet advanced (see Table 2.1). 5 to 6 bits phase shifters are available at microwave frequencies. Both PIN diodes and FETs expand over 1–100GHz band, although, at this high frequency, the loss reaches up to 10dB. The switching time of solid state phase shifters is 1-50ns, depending on the RF power and size of the FETs and PIN diodes used represents an examples on modern linear-phase solid-sate phase shifters. The utilization of microelectromechanical systems (MEMS) technology in solid-state phase shifters has improved their performance significantly as shown in Table 2.2.
Table 2.1: Examples of modern planar solid-state phase shifters.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>Device</th>
<th>Bits</th>
<th>Loss (dB)</th>
<th>Chip Area (mm²)</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>8-12</td>
<td>PIN</td>
<td>4</td>
<td>4</td>
<td>3.7 × 2.3</td>
<td>Wilson et al. [77]</td>
</tr>
<tr>
<td>43-45</td>
<td>FET</td>
<td>3</td>
<td>7.5</td>
<td>2.8 × 2</td>
<td>Aust et al. [78]</td>
</tr>
<tr>
<td>42-46</td>
<td>FET</td>
<td>4</td>
<td>10.5</td>
<td>2.5 × 1.3</td>
<td>Dunn et al. [79]</td>
</tr>
<tr>
<td>61-64</td>
<td>PIN</td>
<td>3</td>
<td>8.8</td>
<td>3.2 × 1.9</td>
<td>Jacomb-Hood et al. [80]</td>
</tr>
</tbody>
</table>

MEMS switched results in lower loss phase shifter, especially at high frequencies 8-100GHz. Additionally, MEMS switched have very small up-state capacitance, that results in wideband performance over similar designs using FET or PIN devices. MEMS phase shifter also leads to significant reduction of DC power in large phased arrays, especially in receive only systems [81] [82] [83].

Table 2.2: Examples of planar MEMS-based phase shifters.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>MEMS Switch Type</th>
<th>Bits</th>
<th>Average Loss (dB)</th>
<th>Chip Area (mm²)</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>7-11</td>
<td>Capacitive</td>
<td>2</td>
<td>1.15 (8GHz)</td>
<td>50</td>
<td>Ratheon [84]</td>
</tr>
<tr>
<td>DC-40</td>
<td>Series</td>
<td>4</td>
<td>2.2 (10GHz)</td>
<td>30</td>
<td>Rockwell [85]</td>
</tr>
<tr>
<td>DC-18</td>
<td>Series (SP4T)</td>
<td>2</td>
<td>0.6 (10GHz)</td>
<td>10</td>
<td>UoM/Rockwell [86]</td>
</tr>
<tr>
<td>26-40</td>
<td>Series</td>
<td>3</td>
<td>2.2 (35GHz)</td>
<td>16</td>
<td>Rockwell [85]</td>
</tr>
</tbody>
</table>
MEMS switches can be fabricated directly on quartz, ceramic, or Teflon-based substrates, therefore, leading to a low-cost implementation of the phased arrays. The only issues with the MEMS switches might be due to their low switching time, in order of 1-30μs that limits their usage to relatively low scanning arrays.

Figure 2.27 represents a 4-bit switched line phase shifter, developed by the University of Michigan in collaboration with Rockwell Corp., with 1.1dB insertion loss at 10GHz and chip area equals 20mm². Recent advances in fabrication technologies and design capabilities have led to reduced size of the phase shifter.

Figure 2.27: Photograph of the Scientific 4-bit MEMS phase shifter using SP4T switches by the University of Michigan/Rockwell [86], ©2002 IEEE.
2.5 Conclusion

In this chapter, we presented the evolution of the tightly coupled array (TCA). As shown, the importance of the TCA principle relates to the fact that creates a frequency independent aperture that supports ultra-wide bandwidth. A practical implementation of that is the connected array. However, the presence of the ground plane limits the bandwidth of the array to only around 2.5:1. A remedy is proposed by Munk [11] who utilized the inter-element capacitance in phased array to control the mutual coupling. Provided for an almost 2x increase of bandwidth (4.5:1) and reduction of array thickness to $\lambda/10$. This unique feature of TCAs paved the way for the low profile, wideband tightly coupled dipole array at Ka-band, presented in chapter 3 of this literature.

We also reviewed various studies about phase shifters for beam forming in phased arrays. Although ferrite based phase shifters are accurate and handle huge power, they are inappropriate for phased array applications due to their huge loss and manual control requirements. On the other hand, solid-state phase shifter utilizing FET and PIN diodes proposes a planar, low profile solution that can be integrated with devices in phased arrays at millimeter-wave and microwave frequencies. Furthermore, it is shown that the loss and size can be further improved by utilizing micro electro-mechanical systems technology in solid-state phase shifter design.
Chapter 3 : A Novel Wideband Phased Array at Ka-Band

3.1 Motivation

Beam agile antenna at a low profile spanning over Ka-band is highly desired for SATCOM and 5G communications at Ka-band. Low profile systems are mandatory for limited budgetary unmanned Ariel vehicles and other compact vehicle mounted applications. The vast distribution of RF systems created the need for costly effective design for many applications. Phased arrays, offer the desired bandwidth at low profile and cost. For agile beam steering that does not squint with frequency, corporate feeding networks is usually used to feed the array. Current phased arrays operate at frequency bands lower than Ka-band [50] [87] [88] [89] [34]. These designs utilized multi-layered PCB structure and convey shorting vias and holes. To date, tightly coupled dipole arrays have never been implemented at Ka-band, creating challenges that need to be addressed. Moreover, current tightly coupled arrays suffer of feeding network discrepancies, such as high loss, bulkiness and complexity of the corporate feeding network. Previous research trials of the tightly coupled arrays have never addressed the feeding network as part of their design. Therefore, a comprehensive study of the array with the feeding network would lead to a design improvement, and a better integration and performance.

In this chapter, a novel coupled dipole array spanning over 18-40GHz that utilizes a single off-the-shelf Rogers 3010 layer is the proposed solution. With simple utilization
of the single-layer dielectric, cost and complexity of the design are reduced significantly. The array is fed through a corporate feeding network that supports agile frequency-independent beam steering. The optimum phase control architecture will be adapted for lowest loss and profile. Accordingly, the array is build using Brick construction, to allow more space, and each unit cell is adapted to accept the phase shifter body on its backside. Furthermore, the design of the unit cell overcomes complexities of previous tightly coupled dipole array designs. It eliminates the need for vias and other fine features. The design is called coupled dipole array (CDA) rather than tightly coupled, as the inter-element capacitance is not fully utilized. The design procedures have relied on a comprehensive parametric analysis of the array unit cell that leads to a better understanding of the input impedance behavior. As an outcome of this study, we found out that the feed gap could be utilized to counteract the inductance of the ground plane and achieve wideband width. We also examined different feeding structures of the unit cell. Alternative proposals for different feeding structures that offer alternate solutions, such as, impedance match, common mode rejections, and compatibility with the integration of the phase shifter at the backside are also examined. Once the final unit cell design with the optimum feed is made, a proof-of-concept finite array is simulated to examine the beam scan performance.

3.2 Parametric Analysis of the Coupled Dipole Array

Reliable analysis and fast design procedures of the coupled phased array requires an in depth understanding of the design parameters and the effect of each parameter on the performance and input impedance of the array. In this section, we analyze the effect of
Figure 3.1: Coupled dipole array over ground plane. (a) Unit cell with geometrical parameters, (b) Equivalent circuit model of the unit cell, and (c) comparison between the input impedance of the unit cell and the circuit model.
each physical parameter of the unit cell on the input impedance of the unit cell itself. The study starts with a conventional unit cell design shown in Figure 3.1 (a). Namely, the cell is made of a linear dipole antenna. The dipole resides quarter wavelength, at the upper frequency of the band, above the ground plane. The cell is smaller than half wavelength in order to avoid grating lobes. Table 3.1 (a) presents the dimensions of the geometry in (mm). Corresponding, an equivalent circuit model of the unit cell is made as illustrated in Figure 3.1 (b). The specifications of the substrate and superstrates in the circuit model are similar to the unit cell design. Whereas, the lumped R, L, and C values were determined based on curve fitting between the simulation and the circuit model. After the input impedance of the unit cell is achieved, it is used as a goal for a curve fitting process, in which the values of the lumped parameters (R, L, and C) are optimized to achieve the best curve fitting as shown in Figure 3.1 (c). The best curve fitting is determined by the

<table>
<thead>
<tr>
<th>Unit Cell Design Figure 3.1(a)</th>
<th>Equivalent Circuit Figure 3.1(b)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( F_{\text{gap}} )</td>
<td>( Z_{\text{sub}} ) 377 ( \Omega )</td>
</tr>
<tr>
<td>( L_{\text{dipole}} )</td>
<td>( Z_{\text{sup}} ) 214 ( \Omega )</td>
</tr>
<tr>
<td>( W_{\text{dipole}} )</td>
<td>( h_{\text{sup}} ) 1.65 mm</td>
</tr>
<tr>
<td>( H_{\text{sup}} )</td>
<td>( h_{\text{sub}} ) 2.5 mm</td>
</tr>
<tr>
<td>( H_{\text{sub}} )</td>
<td>( L ) 0.0 nH</td>
</tr>
<tr>
<td>( L_{\text{cell}} )</td>
<td>( L_{p} ) 0.8 nH</td>
</tr>
<tr>
<td>( W_{\text{cell}} )</td>
<td>( C ) 0.02 pF</td>
</tr>
<tr>
<td></td>
<td>( C_{p} ) 0.009 pF</td>
</tr>
</tbody>
</table>
minimum “relative goal cost” of the optimization method “Random (Global)”, followed by “Random (Local)”). The achieved values of the corresponding optimized circuit are shown in Table 3.1(b).

After the corresponding equivalent circuit model of the unit cell is determined, the effect of each geometrical parameter of the unit cell on its performance, specifically, its input impedance, is observed by closely analyzing a single parameter at a time. The dimension of that parameter of the unit cell is swept over three different values. While the other geometrical parameters of the unit cell are remained unaltered in order to ensure that, the differences between the curves carry the effect of geometrical changes.

After, the three different curves are achieved; an equivalent circuit model of each is implemented through the curve fitting process as already described. In which an optimization process is used to fit the “$Z_{in}$” curve of the circuit model with the one that is obtained from simulation. This results in a new circuit model with new values of the lumped parameters. Therefore, the changes in the lumped components are related to the effect of geometrical changes.

The design parameters that are examined are; the length of the dipole “$L_{dipole}$”: this parameter describes the end-to-end distance between the two dipole arms including the gap of the feed; gap of the feed “$F_{gap}$”; the width of the dipole “$W_{dipole}$”; and the inter-element capacitance “$C_{inter-element}$”. For full wave simulation, and circuit modeling/optimization Ansoft HFSS V15.0 and AWR softwares are used respectively.

Firstly, the dipole length effect is examined. “$L_{dipole}$” is the end-to-end distance between the two arms of the dipole, including the feed gap separation. In order to examine purely
the effect of the “\( L_{\text{dipole}} \)” we maintained the inter-dipole spacing that is responsible for the inter-element capacitance, and the feed gap separation unchanged. Therefore, the element size is changed every time the “\( L_{\text{dipole}} \)” is changed. According to equation (2.2), dimensions of a square unit cell has no contribution on the input impedance value. Therefore, any change in the input impedance value is expected to be due to changes of the “\( L_{\text{dipole}} \)” value. Three different values of the \( L_{\text{dipole}} \) are selected 3.5, 3.75 and 4mm. the middle value, 3.75 represents half-wave length dipole at the largest frequency (40GHz) of the band. Figure 3.2 present the input impedance of the unit cell at the three different scenarios. It is noted from the figure that input impedance become more inductive by increasing the dipole length. Whereas, the real part increases slightly, especially at the higher frequency of the band. This behavior confirms the similarity between the dipole and a straight wire.

![The Effect of Varying “\( L_{\text{dipole}} \)”](image)

**Figure 3.2:** The effect of changing “\( L_{\text{dipole}} \)” on the input impedance of the unit cell.
Figure 3.3: Input impedance of the full wave simulation and the circuit model at three different “L_{dipole}” values; (a) 3.5mm, (b) 3.75mm, and (c) 4mm.
Table 3.2: Optimized circuit values of Figure 3.1(b)

<table>
<thead>
<tr>
<th>Different “$L_{\text{dipole}}$” Values</th>
<th>$L_{\text{dipole}}=3.5\text{mm}$</th>
<th>$L_{\text{dipole}}=3.75\text{mm}$</th>
<th>$L_{\text{dipole}}=4\text{mm}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L$ [nH]</td>
<td>0.4</td>
<td>0.8</td>
<td>1.4</td>
</tr>
<tr>
<td>$L_p$ [nH]</td>
<td>0.35</td>
<td>0.45</td>
<td>0.3</td>
</tr>
<tr>
<td>$C$ [pF]</td>
<td>0.035</td>
<td>0.025</td>
<td>0.03</td>
</tr>
<tr>
<td>$C_p$ [pF]</td>
<td>0.014</td>
<td>0.013</td>
<td>0.013</td>
</tr>
</tbody>
</table>

The achieved input impedance curves are then exported to AWR circuit designer. These curves are used to drive the optimization process of the circuit model, and achieve the lumped R, L, and C values. Figure 3.3 compares between the input impedance of the full wave simulation and the circuit model Figure 3.1 (a) and (b) respectively. As shown, the curves are well fitted. The achieved lumped R, L and C values from the circuit model are summarized in Table 3.2.

From Table 3.2, it is noted that “$L$” increases as “$L_{\text{dipole}}$” increase. Other than this, there are slight unsystematic variations in the other parameters. This confirms that the lumped “$L$” parameter in the circuit model depicts the inductance of the dipole. For better accuracy of the circuit model, one can add a resistor in series that represents the material loss; however, due to the small value of this resistor, it can be neglected.

Secondly, the separation between the dipole arms at the feeding points, “$F_{\text{gap}}$”, is examined. Three different values of the “$F_{\text{gap}}$”, 0.4, 0.6 and 0.8mm are designed. It is noted that changing the “$F_{\text{gap}}$” will affects the “$L_{\text{dipole}}$” and will give ambiguous results. It is important to eliminate the “$L_{\text{dipole}}$” effect, therefore, it is determined to change the size of the unit cell in order to maintain fixed dipole length and to eliminate the effect of...
Figure 3.4: The effect of changing “$F_{gap}$” on the; (a) input impedance of the unit cell, (b) VSWR.
Figure 3.5: Input impedance of the full wave simulation and the circuit model at three different "F<sub>gap</sub>" values; (a) 0.4mm, (b) 0.6mm, and (c) 0.8mm.
“$L_{dipole}$”. Figure 3.4 (a) presents the input impedance of the unit cell at the three different scenarios. As shown by the figure, increasing “$F_{gap}$”, leads to an increase in the imaginary part of the input impedance, leading to a wider bandwidth as shown in Figure 3.4 (b). As noted, the feed-gap has a significant effect on the input impedance. It can be used to cancel/reduce the inductance of the ground plane replicating the same effects as the inter-element capacitance leading to the wider bandwidth shown in Figure 3.4 (b) Next, the simulated input impedance curves are used to drive the circuit model, leading to new values of the lumped components. These new values reflect the changes in “$F_{gap}$”. Figure 3.5 compares between the input impedance of the circuit model that is achieved by curve fitting, and the original curves of the full wave simulation. The optimization process led to a very precise and accurate curve fit as shown indicating that indeed the circuit model is very reliable, and the values of the lumped components, shown in Table 3.3, are accurate. By looking at the values in Table 3.3, increasing “$F_{gap}$”; reduces “$C_p$”; and increases “$L_p$”. Therefore, the lumped “$L_p$” and “$C_p$” are a good modelling representation of the feeding gap “$F_{gap}$”. 

Table 3.3: Optimized circuit values of Figure 3.1(b) 

<table>
<thead>
<tr>
<th>Different “$F_{gap}$” Values</th>
<th>$F_{gap} =0.4\text{mm}$</th>
<th>$F_{gap} =0.6\text{mm}$</th>
<th>$F_{gap} =0.8\text{mm}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L$ [nH]</td>
<td>0.85</td>
<td>0.8</td>
<td>0.84</td>
</tr>
<tr>
<td>$L_p$ [nH]</td>
<td>0.36</td>
<td>0.46</td>
<td>0.55</td>
</tr>
<tr>
<td>$C$ [pF]</td>
<td>0.028</td>
<td>0.024</td>
<td>0.023</td>
</tr>
<tr>
<td>$C_p$ [pF]</td>
<td>0.015</td>
<td>0.013</td>
<td>0.01</td>
</tr>
</tbody>
</table>
Thirdly, the inter-element capacitance “$C_{\text{inter-element}}$” is swept over 0.01, 0.1, and 0.2 values. The corresponding changes of the input impedance are shown in Figure 3.6.

![The Effect of Varying “$C_{\text{inter-element}}$”](image)

Figure 3.6: The effect of changing “$C_{\text{inter-element}}$” on the input impedance of the unit cell.

From the figure, as “$C_{\text{inter-element}}$” decreases, the input impedance at the lower band becomes more inductive. Similarly, the real value of the input impedance increases.

The circuit model is achieved through the curve fitting. As shown in Figure 3.7 high match between the circuit model and the full wave simulation is achieved. In addition, the optimized values of the lumped parameter, presented in Table 3.4, indicates that “$C$” decreases as “$C_{\text{inter-element}}$” increases. Furthermore, we can notice that “$C_{\text{inter-element}}$” contributes to the value of “$L_p$”. As “$C_{\text{inter-element}}$” increases, “$L_p$” decreases.
Figure 3.7: Input impedance of the full wave simulation and the circuit model at three different “$C_{inter-element}$” values; (a) 0.01mm, (b) 0.1mm, and (c) 0.2mm.

(a) $C_{inter-element} = 0.01 \text{mm}$

(b) $C_{inter-element} = 0.1 \text{mm}$

(c) $C_{inter-element} = 0.2 \text{mm}$
Table 3.4: Optimized circuit values of Figure 3.1(b)

<table>
<thead>
<tr>
<th>Different “C&lt;sub&gt;inter-element&lt;/sub&gt;” Values</th>
<th>C&lt;sub&gt;inter-element&lt;/sub&gt; = 0.01 mm</th>
<th>C&lt;sub&gt;inter-element&lt;/sub&gt; = 0.1 mm</th>
<th>C&lt;sub&gt;inter-element&lt;/sub&gt; = 0.2 mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>L [nH]</td>
<td>0.85</td>
<td>0.58</td>
<td>0.87</td>
</tr>
<tr>
<td>L&lt;sub&gt;p&lt;/sub&gt; [nH]</td>
<td>0.45</td>
<td>0.42</td>
<td>0.32</td>
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<tr>
<td>C [pF]</td>
<td>0.023</td>
<td>0.016</td>
<td>0.013</td>
</tr>
<tr>
<td>C&lt;sub&gt;p&lt;/sub&gt; [pF]</td>
<td>0.013</td>
<td>0.013</td>
<td>0.013</td>
</tr>
</tbody>
</table>

Finally, the width of the dipole “W<sub>dipole</sub>” effect is examined. “W<sub>dipole</sub>” is swept over 0.2, 0.5 and 0.7 mm. It is commonly known that the dipole width affects its input resistance. As shown in Figure 3.8, the resistance of the input impedance increases as “W<sub>dipole</sub>” decreases. Whereas, at the low frequencies, the imaginary part become more inductive as

Figure 3.8: The effect of changing “W<sub>dipole</sub>” on the input impedance of the unit cell.
Figure 3.9: Input impedance of the full wave simulation and the circuit model at three different \( W_{\text{dipole}} \) values; (a) 0.2mm, (b) 0.5mm, and (c) 0.7mm.
“$W_{dipole}$” increases. Similarly, for the real part as expected when the inter-element capacitance increases as “$W_{dipole}$” increases leading to a higher inductive contribution at the low frequencies as shown earlier (in “$C_{inter-element}$”-examination scenario). Furthermore, as “$W_{dipole}$” increases, the fed gap changes. By examining the behavior of the “$L_p$” and “$C_p$” and comparing it with its behavior in case of “$F_{gap}$”-examination, it can be observed that increasing the dipole width “$W_{dipole}$” has similar effect as reducing the feed gap separation “$F_{gap}$”. As “$W_{dipole}$” increases, the feed capacitance “$C_p$” decreases. This case is similar to reducing “$F_{gap}$” that led to an increase in “$C_p$” value as well. Similarly, the feed inductance “$L_p$” and “$W_{dipole}$” also have an inverse relationship; therefore, the dimensions of the feed gap area are significant in controlling the input impedance of the dipole, and both the gap width and separation give collaborative effect. Wider gap width can replace tighter feed gap separation “$F_{gap}$” and vice versa. This phenomenon explains the higher bandwidth of the bowtie over the dipole. Since Bowtie is narrower than the dipole at the feeding terminals, this means tighter “$F_{gap}$”, and wider bandwidth performance. The maximum bandwidth that can be achieved by coupled array

<table>
<thead>
<tr>
<th>Different “$W_{dipole}$” Values</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
</tr>
<tr>
<td>$L$ [nH]</td>
</tr>
<tr>
<td>$L_p$ [nH]</td>
</tr>
<tr>
<td>$C$ [pF]</td>
</tr>
<tr>
<td>$C_p$ [pF]</td>
</tr>
</tbody>
</table>

90
utilizing “$F_{\text{gap}}$”, is equal approximately 2.8:1 at VSWR<2. “$F_{\text{gap}}$” adds significant capability to wideband array design, especially at low profile. In low profile arrays at high frequency band, it is difficult to achieve large inter-element capacitance, therefore, “$F_{\text{gap}}$” can be used in collaboration with the inter-element capacitance to increase the bandwidth.

Similarly, the effects of the substrate and superstrate loadings on the input impedance of the unit cell are examined. The substrate and superstrate layers are represented by transmission line sections with properties determined by the propagating Floquet mode with in each layer. According to equation (2.6) we should know the effective dielectric constant of the material that fills the waveguide in order to determine its impedance. The challenge arise when the dielectric material is not filling the waveguide entirely as in the unit cell shown in Figure 3.10. In this case, the effective dielectric constant should be

Figure 3.10: Coupled dipole array unit cell residing on a vertical dielectric layer that simplifies material loading.
determined in order to calculate the waveguide impedance. In previous designs, vertical
dielectric layers with low dielectric contact (~2) were used [50] [90]. The effective
dielectric constant of the unit cell is assumed unaltered by the dielectric layer, therefore,
it was equal to air. However, if the dielectric constant is high (~10), the more optimal
step is to determine the new effective dielectric constant for better accuracy.

Firstly, we examined the substrate alone while superstrate is assumed air as shown in
Figure 3.11 (a). Rogers3010 substrate with “\(H_{\text{sub}}\)”=2.5mm height and \(\varepsilon_r\)≈10 is used. The
input impedance of the unit cell is determined using the full wave simulation. Then, it is
used to drive circuit model to find the optimum value of the relative dielectric constant
(\(\varepsilon_{r,\text{eff}}\)) of the substrate. The lumped R, L, and C values are determined earlier and
assigned to the circuit model. Optimum \(\varepsilon_{r,\text{eff}}\) that gave the lowest “relative goal cost” of

![Diagram](image1)

![Diagram](image2)

Figure 3.11: (a) Coupled array unit cell with Rogers 3010 substrate loading, (b) input
impedance at broadside scan.
the optimization process at broadside, $E$- and $H$-planes is $\varepsilon_{r_{\text{eff}}}=1.6$. Figure 3.11(b) shows the match between the circuit model and full wave simulation at broadside scan scenario. Similarly, the superstrate is examined. Geometry is designed, as shown in Figure 3.12 (a), in which Rogers 3010 is assigned to the superstrate while the substrate is assumed air. Superstrate height “$H_{\text{sup}}$”=1.6mm, whereas, other geometrical parameters remain as in previous scenario. The obtained input impedance is used to drive the circuit model and find the $\varepsilon_{r_{\text{eff}}}$ of the superstrate. Figure 3.12 (b) shows the match between the simulation and the circuit model. As shown, the two curves are well matched at low frequencies, whereas, the match degrades as frequency increases. The achieved $\varepsilon_{r_{\text{eff}}}$ after optimizing the circuit model was $\varepsilon_{r_{\text{eff}}} \approx 1.5$. This value is confirmed with the one obtained earlier.

Figure 3.12: (a) Coupled array unit cell with Rogers 3010 substrate loading, (b) input impedance at broadside scan.
Accordingly, we can generalize that effective relative permittivity $\varepsilon_{r_{\text{eff}}}$ of a non-homogenous region that consists of air with other dielectric material, can be approximated by the following relation:

$$\varepsilon_{r_{\text{eff}}} = \frac{(y-x)\varepsilon_o + x^2\varepsilon_r}{y}$$

(3.1)

Where, $x$ is the thickness of the dielectric, and $y$ is the thickness of the waveguide as shown in Figure 3.13.

![Figure 3.13: Air-filled region that includes another dielectric material with relative permittivity equals $\varepsilon_r$.](image)

Finally, The effect of aperture size (grid size) of the unit cell is examined. In order to generalize our results, we designed a new array structure. In this structure we used a bowtie radiator instead of a dipole. Furthermore, we utilized over-lapping inter-element capacitance as shown in Figure 3.14. The simulation is run up to a frequency equals double the resonance frequency of the dipole ($\lambda = L_{dipole}$). Three different unit cell sizes,
3mm, 1.5mm, and 0.75mm, are simulated. By looking at the corresponding impedance plots, it is noticed that the real and imaginary parts behaves similarly in all scenarios. Except at 3mm cell size and specifically at 50GHz, the radiation resistance is higher as the element reaches its resonance. Also, the capacitance increases as the cell size decreases, that is referred to the reduction in the inter-element capacitance value. Overall, no big benefits are gained by reducing the grid size, that is worth over sampling the array.

Figure 3.14: Impedance of dipole arrays at three different unit cell widths; 3mm, 1.5mm, and 0.75mm.
and increasing the cost and complexity of the design. At the same time, the unit size should not be increased up to a limit where grating lobes start to appear. Thus, a good compromise to the unit cell width would be right below $\frac{\lambda_{\text{high}}}{2}$ for 45° scan, where $\lambda_{\text{high}}$ is the wavelength at the highest frequency.

From the previous studies we conclude the following:

- The input impedance is effected by the co-linear separation between the dipole elements in the array that is represented by “$L_{\text{dipole}}/L_{\text{cell}}$”.
- The dipole inside an array environment behaves different from its behavior alone, and, the input impedance of a dipole in coupled array is not related to its electrical length.
- The size of the unit cell has a negligible effect on the input impedance as soon as the ratio between the lengths over the width is unity. Whereas, for non-square unit cell, the ratio between the width “$W_{\text{cell}}$” and the length “$L_{\text{cell}}$” of the cell, controls the input impedance as described by equation (2.2).
- The length of the dipole has a negligible effect on the input impedance.

More details about the effect of the physical features of the unit cell on the real and imaginary parts of the input impedance are presented in Table 3.6.
After the unit cell design is determined, it is desired to integrate the appropriate feeding structure to the unit cell. This feeding structure has to be impedancily matched with the unit cell, and it has to support the same wideband as the unit cell does. The appropriate feeding structure should be compatible with the unit cell as well. On the next section, a survey of different feeding structures is proposed and the optimum design will be selected and integrated with the unit cell.

### 3.3 Wideband Feeding Structure for the Coupled Dipole Array Unit Cell

In previous sections, it was determined that the size of the beam forming system in phased arrays can be minimized by inserting the phase shifter at the unit cell element above the ground plane. For this purpose, brick construction of the array provides the desired space for the phase shifter. Therefore, the radiator unit element has already been

<table>
<thead>
<tr>
<th>Parameter</th>
<th>( \text{Im } { \text{Zin} } )</th>
<th>( \text{Re } { \text{Zin} } )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( H_{\text{sub/dipole}} )</td>
<td>Inversely Proportional at Low Frequency</td>
<td>Proportional at Low Frequency</td>
</tr>
<tr>
<td>( W_{\text{Cell}} = L_{\text{Cell}} ) (Fixed Inter-element Capacitance)</td>
<td>No Systematic Effect. The effect is referred to change in dipole electrical length.</td>
<td>No Systematic Effect. The effect is referred to change in dipole electrical length.</td>
</tr>
<tr>
<td>( W_{\text{Cell}} = L_{\text{Cell}} ) (Fixed Dipole Length)</td>
<td>Inversely Proportional</td>
<td>Inversely Proportional</td>
</tr>
<tr>
<td>( W_{\text{dipole}} )</td>
<td>Proportional at Low Frequency</td>
<td>Proportional at Low Frequency</td>
</tr>
<tr>
<td>( L_{\text{dipole}} ) (Fixed Cell Dimensions)</td>
<td>Proportional</td>
<td>Proportional</td>
</tr>
<tr>
<td>( F_{\text{gap}} )</td>
<td>Proportional</td>
<td>Proportional</td>
</tr>
</tbody>
</table>
designed on a planar vertically aligned dielectric layer. Similarly, the appropriate feeding of the unit cell is to be designed and integrated on this structure.

One common design challenges for wideband phase arrays is the design of a feeding structure that does not deteriorate the natural bandwidth behavior of the array. In general, wideband arrays utilize radiating elements that use balanced feeds. In balanced line, there are three conductors, where two of them are similar, and balanced with respect to the third conductor that simplifies the ground. This is in contrast with the unbalanced feed, where the feed nodes are fed by signal and ground. Balanced lines support two modes, common mode, and differential mode. In common mode excitation, the conductors are fed with exact same signals with no phase delays. Whereas, in differential excitation mode, the two lines are fed with 180° phase difference. Using common mode excitation, interference from other circuits can be rejected. On the other hand, by using differential signals, we can prevent radiation of the leakage-signal.

From the antenna prospective, common mode propagation is undesired in transmission lines. If the feeding lines support common modes, they radiate and distort the main radiation patter. Furthermore, they increase cross-polarization level, degrades the bandwidth and gain [59]. Even when exciting differential mode, it is possible that common mode appears at the feeding lines. They can appear due to an incident wave, especially, when the feed lines are left unshielded. This resonance can be solved through the utilization of a “feed organizer” [59] [91] that shields the feed lines and reduces the electrical distance between neighboring feed lines.
However, the issues with the balanced feeds is that they are usually driven by electronics and transmission lines that typically employ unbalanced feed lines, such as, microstrip, stripline and coaxial cables. Therefore, a transformation between unbalanced to balanced is desired. This can be done through different approaches; firstly, using a transitional structure that transforms between the different types of transmission lines, i.e. CPW or strip to coupled transmission line; secondly, by using a device called balun; or thirdly, by using shorting posts to eliminated the common mode resonance [92].

After rigid and in-depth due diligence in our observations, we present several nominated planar feeding structure as shown in Figure 3.15. These feeding structures are compatible with the integrations of the phase shifter, also, they provide differential current to the antenna that is desired to avoid the generation of the common mode. The designs in a), b) and e) are dipole antennas that are fed with balun. As seen, the balun transforms the unbalanced microstrip line into balanced coupled line that is feeding the dipole. On the other hand, the figure in c) and d) shows dipoles that are fed by coplanar transmission (CPW) lines. In these design, a transitional structure is used to transit between the unbalanced input feeding line and the balanced line that is connected to the feeding port of the antenna. Compared with strip/microstrip lines, CPW offers less dependency on the substrate, and much lower material loss (less than one-fourth) of the microstrip as shown in Table 3.7. Moreover, the CPW fed design in d) and c) requires metallization on one side of the dielectric layer. This reduces the fabrication cost of a large array. Accordingly, the CPW line simplifies a good feeding option to start with.
Figure 3.15: Different feeding structures; a) Dipole antenna with integrated balun [93], ©2015 IEEE, b) Wideband and compact quasi-Yagi antenna integrated with balun of microstrip to slotline transitions [94], Reproduced by permission of the Institute of Engineering & Technology, c) wideband Sierpinski fractal bow-tie antenna [95], ©2006 IEEE, d) Printed Antenna Fed by Coplanar Waveguide [96], ©2014 IEEE, and e) Marchand-balun Fed Dipole antenna [97], ©2008 IEEE.

Table 3.7: Transmission lines comparison.

<table>
<thead>
<tr>
<th>Type</th>
<th>Effective Dielectric Constant ($\varepsilon_r$)</th>
<th>Loss(dB)/mm @24GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Coupled Microstrip</td>
<td>$\approx 6$</td>
<td>$\approx 0.03$dB</td>
</tr>
<tr>
<td>Microstrip</td>
<td>$\approx 6$</td>
<td>$\approx 0.1$dB</td>
</tr>
<tr>
<td>Coplanar Waveguide</td>
<td>$\approx 3.3$</td>
<td>$\approx 0.018$dB</td>
</tr>
<tr>
<td>Grounded Coplanar Waveguide</td>
<td>$\approx 5.9$</td>
<td>$\approx 0.07$dB</td>
</tr>
</tbody>
</table>

3.4 Low-Loss Coupled Dipole Array

The dipole radiator is printed on a 254μm thick Rogers 3010 dielectric layer. Bowtie dipole shape is used with high feed gap for widest bandwidth, as shown in Figure 3.16.

Further distance of separation is not recommended, as the impedance behavior of the dipole distracts significantly, other modes appear and the dipole no longer acts as a radiator. The dipoles are printed on a single side of the dielectric layer, as a result, no
overlapping capacitance is created. The inter-element tips are separated by 75μm to satisfy the fabrications limitations of the PCB fabrication technology. After the optimum dimensions are determined, the dipole is fed through CPW transmission lines. Two parallel lines, with a characteristic impedance of 75Ω, where used with 180° phase difference to provide differential balanced feeding as shown in Figure 3.16. The CPW lines are connected through a common ground layer in the middle. Open stub is used in parallel with the CPW lines to improve the impedance match. The open stub is made of a slotline transmission line that is compatible with the structure and made of a single metallic layer. The optimum slotline is found to be 0.6mm as show in Figure 3.17 where approximately 2:1 BW (17-34GHz) is achieved. The performance of the optimum unit
Figure 3.16: Impedance bandwidth at three different lengths of the short stub.

cell design shown in Figure 3.16 is presented in Figure 3.18. In (a), an impedance bandwidth spanning over 17-34GHz is achieved up to 45° scan in both E- and H-planes. Radiation efficiency of the design exceeds 85% as shown in (b). This proves the low material loss and high impedance match. Furthermore, the achieved cross-polarization shown in (c), is greater than 17dB.

After the optimum cell is achieved, the efficient procedure to follow up is to upgrade the unit cell for the integrations of the phase shifter. The phase shifter to be used is a solid-state phase shifter with MEMS technology for switching and capacitors. More detail of the phase shifter will be presented in chapter 4. The dimensions of the phase shifter are 1.8×1.8×0.35mm. For our unit cell, it is desired to use two phase shifters one on each CPW line. The phase shifters will be sided together. Since, the phase shifter consists of microstrip transmission lines, one should create a transition from the CPW to microstrip.
Figure 3.18: The Optimum unit cell performance at broadside, 45° E-plane scan, and 45° H-plane scan; (a) VSWR, (b) radiation efficiency, and (c) Polarization purity.

Furthermore, the phase shifter is to be integrated at the backside of the unit cell; therefore, having a ground plane for the phase shifter at the backside of the unit cell is more effective. After adding the ground plane, a band gap appears in the operational bandwidth. To examine the reason of this band gap, the ground plane height is varied as shown in Figure 3.19 (a), and the corresponding behavior of the impedance bandwidth is recorded as in Figure 3.19 (b).
Figure 3.19: Ground plane effect on the input impedance of the unit cell. (a) ground plane at different height, and (b) the corresponding VSWR of each scenario.

As noted, the bandgap appears at frequency, where the dipole is far from the ground plane by \( \lambda/4 \). The radiated field from the antenna reflects back from the ground and adds destructively (out of phase) with the original radiated field, as a result of that, the dipole height above the vertical ground plane is increased. The final unit cell design that is adapted with the integration of the phase shifter is shown in Figure 3.20.
Figure 3.20: CDA unit cell that is adapted for the integration of the phase shifter at its backside.

The CPW transmission line is transit to the backside and transform into microstrip that is compatible with the phase shifter. Ideal delay lines are used at this point. In addition, the phase shifter is simplified by a silicon layer of 1.8×1.8×0.35mm dimensions. Two sided phase shifter are desired per each unit cell. As noted, the dipole height increases to 4.4mm to compensate for the effect of the vertical ground plane. The performance of the unit cell is presented in Figure 3.21.
Figure 3.21: Unit cell performance; (a) VSWR at different scan angles of the H-plane, (b) VSWR at different scan angles of the E-plane, (c) Total Efficiency, and (d) polarization purity.

An impedance bandwidth ~2:1 is achieved during scan ±45° in both E- and H-planes. Overall efficiency, including mismatch loss, is greater than 50%. In addition, the polarization purity is better than 20dB. Unfortunately, the bandwidth of the design is limited and it cannot be increased further. The two sided phase shifters restrict the minimum cell width to 3.6mm. Furthermore, the inter-element capacitance is too minuscule and cannot be increased due to PCB fabrication requirements (minimum spacing between to metallic shapes should be >75μm). The design suffers from several
limitations. The most significant being complexity, which includes transition from CPW to microstrip, vias, and transition from front to back; cost, as each unit cell required two phase shifters; and limited bandwidth, maximum achieved bandwidth is ~2:1.

Back to Figure 3.15, further examination of designs a), b) and e) show these designs consist of dipole radiator that is fed through a planar balun. A balun is a device that performs transformation between balanced to unbalanced feed. It provides the desired differential signals at the feeding point of the dipole. It can be made of different kinds of transmission lines. The ones shown in Figure 3.15 are made of microstrip lines. Therefore, the phase shifter can be seamlessly integrated on them and no transitional structure is desired. Furthermore, the topology of feeding consists of one input line, thus, it is only required to have one phase shifter per unit cell, the thing that reduce the complexity, cost of the design significantly, and it allow for much smaller unit cell.

3.5 Ultra-Wideband Coupled Dipole Array

A balun that is used in wideband tightly coupled dipole array should possess the following features:

- Excite the differential mode.
- Transform the impedance between the feeding cable and the dipole
- Maintain scanning capability.
- Supports the operational bandwidth of the array.

Designing a balun that combines all these features can be a major technical challenge. Many successful designs have been proposed utilizing wideband balun as a feeding to
their array [34] [50]. Additionally, in some designs, the balun is also used as a matching network to improve the performance of the array [50]. In the following section, the marchand balun is described in more details, and a marchand-balun fed CDA is proposed.

3.5.1 Wideband Marchand-Balun Feeding Structure

There are two types of balun: coiled transformer type and transmission line type. Coil transformer type consists of primary and secondary coils that are coupled through the magnetic flux. This type has several advantages such as its low profile, wide bandwidth and varied impedance transformation ratios; however, coiled baluns are expensive. On the other hand, transmission line baluns, which are based on transmission lines are more compact, less expensive and more importantly, easier to integrate and fabricate with monolithic circuit designs. Due to the nature of electromagnetic coupling in transmission line baluns, they can be designed more accurately than the coiled ones. Transmission line baluns usually incorporates several lines and operate based on delay lines and open/short stubs [98]. Some of the common examples are the Guanella balun [99], the marchand and the double-Y baluns [100].

The Marchand balun is a wideband design that was first introduced in 1944 [101]. Consisting over a decade of bandwidth [102] [103] [104], it can be readily integrated with dipole-like elements [105]; therefore, it is ideal for the tightly coupled dipole array. Furthermore, it can be integrated with monolithic circuits such as LNA and phase shifter. The most basic marchand-balun design is described in Figure 3.22 (a). The balun consists of a coaxial that is contained in a conducting box. The equivalent circuit that represents
the marchand-balun [106] [102] is also presented in Figure 3.22 (b). As shown, the balun reaches common mode suppression over wideband through symmetry. The balanced output currents are shown with the undesired common mode currents. As shown, the common mode currents interact destructively at the center of the balun, and unbalanced currents propagate to the balanced lines. The cancellation process is frequency independent and it operates over wide range of frequency. Therefore, the only remaining challenge in the design of the balun is wideband impedance matching.

![Diagram of balun structure](image)

Figure 3.22: (a) Original Marchand-Balun structure as proposed in [101], and (b) equivalent circuit as given in [106] [102].

Figure 3.23 presents the folded marchand balun [107] explaining the current flow in it. Similar to the original marchand balun, the folded type exploits symmetry to excite purely differential mode. As demonstrated, the undesired common mode excitation is subject to current cancellation at the junction of the two lines. In Figure 3.23 (b) the
equivalent circuit of the folded marchand balun is presented. The main components are
the unbalanced transmission line with characteristic impedance of $Z_f$, open stub
transmission line and shorted transmission line with characteristic impedances equals $Z_{oc}$
and $Z_{sc}$ respectively.

![Diagram of folded Marchand balun](image)

Figure 3.23: (a) The folded Marchand balun as described in [107], (b) the corresponding
equivalent circuit of the folded-Marchand balun.

Early Marchand balun designs were mainly designed with coaxial cables. More recent
versions are designed in a variety of MMIC-based [102] and PCB-based transmission
lines [106] [103] [108] [109] [110]. This allows integration with microwave devices and
antenna. In the next section, a PCB microstrip line based folded Marchand balun is
integrated with the tightly coupled dipole array unit cell to provide its required balanced
feeding.

### 3.5.2 Marchand-Balun Fed Coupled Dipole Array

One of the most common design challenges for wideband phase arrays is the design of a
feeding structure that does not deteriorate the natural bandwidth behavior of the array.
Ideally, the balun can be integrated with each array elements, and this integration can be performed on the top of ground plane to save weight and size. Furthermore, one can tune the reactance of the balun to cancel that of the array, as is usually done in Vivaldi arrays [111]. That is, the balun may be viewed as part of the impedance matching network for the array. By utilizing the balun as part of the matching network of the array, a higher-order impedance match can be obtained, leading to a higher bandwidth as proposed by Fano in [112]. Figure 3.24 presents a version of the marchand-balun fed coupled dipole array (MB-CDA). As shown in (b) short stub can be used as well, instead of the open stub. In this case, the stub length should be too small (~0) to give equivalent impedance.

Figure 3.24: (a) Unit cell design of the MB-CDA, (b) Marchand balun constructed from λ/4 open stub.
A Marchand balun constructed from strip transmission lines is both compact and theoretically capable of operating over 10:1 bandwidth. The equivalent circuit in Figure 3.25 represents a superstrate-loaded array with 3rd order matching network of the Marchand balun. Because the superstrate and substrate themselves simplifies two additional stages, the total order of the array is n=5. The matching circuit is optimized for maximum bandwidth by using AWR Microwave Office tool as shown in Table 3.8. The achieved bandwidth (≈8.4:1, VSWR<2), as shown in Figure 3.26, seems to be consistent with the theoretical Fano limit of 5th order network presented in [113]. In Table 3.8, it is demonstrated that extreme impedance ratios, $Z_{OC}<<Z_{SC}$, are desired by a single stage balun to operate over wide bandwidth [104]. In our case we incorporated a folded Marchand balun, that fulfill the fabrication limitations, with the CDA. The optimum resulting structure for wideband matching is shown in Figure 3.27 (a).

Figure 3.25: CDA equivalent circuit with a Marchand balun feed, that simplifies 5th order feed matching network.
Table 3.8: Optimized circuit values of CDA with integrated balun

<table>
<thead>
<tr>
<th>Equivalent Circuit Figure 3.25</th>
<th></th>
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<tr>
<td>$Z_{0}$, $Z_{sub}$</td>
<td>377 Ω</td>
</tr>
<tr>
<td>$Z_{sup}$</td>
<td>214 Ω</td>
</tr>
<tr>
<td>$Z_{DC}$</td>
<td>1 Ω</td>
</tr>
<tr>
<td>$Z_{SC}$</td>
<td>300 Ω</td>
</tr>
<tr>
<td>$Z_{TCDA}, Z_{d}$</td>
<td>135 Ω</td>
</tr>
<tr>
<td>$h_{sub}$</td>
<td>2.9 mm</td>
</tr>
<tr>
<td>$h_{sup}$</td>
<td>2.1 mm</td>
</tr>
<tr>
<td>$h_{OC}$</td>
<td>2.3mm</td>
</tr>
<tr>
<td>$h_{SC}$</td>
<td>2 mm</td>
</tr>
<tr>
<td>$L_{dipole}$</td>
<td>0.45 nH</td>
</tr>
<tr>
<td>$L_{feed}$</td>
<td>0.18 nH</td>
</tr>
<tr>
<td>$C_{coupling}$</td>
<td>0.1 pF</td>
</tr>
<tr>
<td>$C_{feed}$</td>
<td>0.03 pF</td>
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</table>

Figure 3.26: Maximum optimized bandwidth of the equivalent circuit in Figure 3.25.
Figure 3.27: (a) MB-CDA optimum unit cell design, (b) the corresponding impedance bandwidth at broadside and beam scan scenarios.
Figure 3.28: Input impedance comparison between the full wave simulation and the circuit model; (a) broadside, (b) 45° Scan in E-plane, and (c) 45° Scan in H-plane.
A maximum bandwidth of 2.4:1 is achieved (17-40GHz) VSWR<2, as illustrated by Figure 3.27 (b), after the utilization of the marchand balun as a matching network. The match between the circuit and the full wave simulation at three different scan scenarios, namely, broadside, 45° scan in E- and H-planes are shown in Figure 3.28. Furthermore, the corresponding lumped values of the equivalent circuit model in Figure 3.25 are presented in Table 3.9.

Table 3.9: Optimized circuit values of MB-CDA

<table>
<thead>
<tr>
<th>Value Parameter</th>
<th>Equivalent Circuit Figure 3.25</th>
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<tr>
<td>$Z_{sup}$</td>
<td>238Ω</td>
</tr>
<tr>
<td>$Z_{sub}$</td>
<td>238Ω</td>
</tr>
<tr>
<td>$h_{sup}$</td>
<td>1.7mm</td>
</tr>
<tr>
<td>$h_{sub}$</td>
<td>~2.2mm</td>
</tr>
<tr>
<td>$C_{coupling}$</td>
<td>~0.5pF</td>
</tr>
<tr>
<td>$L_{dipole}$</td>
<td>~0.6nH</td>
</tr>
<tr>
<td>$C_{feed}$</td>
<td>0.06pF</td>
</tr>
<tr>
<td>$L_{feed}$</td>
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</tr>
<tr>
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</tr>
<tr>
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<tr>
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</tr>
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<td>~75Ω</td>
</tr>
<tr>
<td>$Z_{TCDA}$</td>
<td>75Ω</td>
</tr>
</tbody>
</table>
3.5.3 Unit Cell with Integrated Phase Shifter

After the optimum MB-CDA design is reached, the design is updated to host the phase shifter. As mentioned, the phase shifter to be used is a solid-state phase shifter with MEMS technology in it, as will be described in chapter IV. The circuitry of the phase shifter resides on a silicon die of 1.8×1.8×0.35mm size. Accordingly, we decided to integrate the phase shifter at the marchand balun utilizing the same microstrip line. Therefore, this reduces the complexity as the need for transitional structure is eliminated. Also, the same ground plane of the marchand balun is used for the phase shifter as well, as shown in Figure 3.29 (a). This simplifies the design again, it also eliminates the need for additional ground plane for the phase shifter. Notice, the dipole height above the ground plane remains almost unchanged, except for slight increase to improve the impedance match. In other words, there was no need to increase the dipole height after adding the ground plane of the phase shifter, as we did in CPW-CDA in Figure 3.20. Simply because the radiated field from the dipole and the reflected field from the ground, although it is small, are added constructively. Finally, the balun at the back is transit to the phase shifter through a small via. The via with the circular cut in the metal acts as a coaxial. Therefore, the correct design dimensions should be approached to guarantee the smooth transmission of the signal through that via. Practically, that hollow can be empty, and it should be crammed with a dielectric. The achieved impedance bandwidth of the MB-CDA design with phase shifter is presented in Figure 3.29 (b), and 3.2:1 BW (13-42GHz) at VSWR<2.5 are achieved over 45° scan in both E- and H-planes.
Figure 3.29: Updated MB-CDA design for hosting the phase shifter, and (b) impedance bandwidth at broadside, and 45° scan in both $E$- and $H$-planes.
3.6 Simulated Finite Array performance

Most of the work so far is based on the unit cell design. This unit cell represents the performance of an infinite array. However, in reality, infinite array cannot be achieved, thus it has to be truncated to a limited number of elements. This truncation of the array is found to increase the reflection at the feeding ports, due to what is called surface waves. Surface waves travel back and forth at the aperture of the array, and alter the induced voltages at the elements, leading to variations in the active impedance values and impedance mismatch. Figure 3.30 represents a 4×4 tightly coupled dipole array and its VSWR when compared to its infinite geometry. As shown in (b), the VSWR has increase after the truncation of the array into finite size. To solve this issue, several techniques were proposed, such as, taper excitation, most importantly, the characteristic mode excitation [114]. This technique is based on non-uniform excitation of the elements to compensate for the induced voltages that are added by the surface waves. Thus, better match will be achieved. The characteristic mode excitations is applied and found to give perfect matching for array elements, leading to a high efficiency [115]. However, the main drawback of this technique is that it requires a custom-made feeding network. The second approach is to attenuate the surface waves in their way back from the edges to the center elements. This can be accomplished through edge element termination technique. That utilizes resistors at the edges to eliminate the surface waves. Since this technique is risky, it can only be used in applications where the efficiency is not a critical system requirement. Different termination types can also be used, such as short circuit, open circuit or a combination of them. In this section, the optimum termination type, and
Figure 3.30: (a) Finite 4×4 Array and (b) VSWR comparison between the finite and Infinite array.

the number of terminated elements will be determined based on VSWR, and radiation efficiency. The first step is to determine the active VSWR for all the active elements of a given finite array with N-ports. One way is to simulate the array every time the termination changes, by using full wave simulation tools, nevertheless, this might be time
Another way, is to run a single simulations and to obtain the N x N mutual impedance matrix of the full N-port array. Then impose any kind of termination by simple post-processing calculations. From the mutual impedance, one can calculate the active impedance matrix or the scattering matrix, where both can lead to the active VSWR.

In Figure 3.31 (a), an 8x8 array is simulated were all elements are fed equally with ~200Ω feed line. Then, columns 1st and 8th are terminated by 200Ω resistor, where all other elements are fed by 200 Ω feed lines. As shown in Figure 3.32 (b), the VSWR values at the feed lines are lower than in the fully excited array. As a result, one can conclude that edge termination with a resistance, lowers the VSWR of the active elements. In addition, the corporate network VSWR remains low after imposing the termination. The corporate network, for simplicity, assumes 1:N divider/combiner, that is lossless, perfectly matched at all ports, with perfect phase balance and infinite isolation between the output ports. In Figure 3.32(c), we terminated 1st and 8th rows with 200 Ω
Figure 3.32: Active VSWRs of the array depicted in Figure 3.31 at different row/column edge terminations given on top of the figures.
resistor, where all other elements are fed by 200 Ω lines. As shown, the VSWR remains at about the same level as in fully excited array, indicating that row termination have no effect on the performance of the array. In Figure 3.32 (d) columns 1ˢᵗ, 2ⁿᵈ, 7ᵗʰ and 8ᵗʰ are terminated by a 200 Ω resistor, whereas, the other elements are excited by a 200 Ω lines. The result shows that terminating two columns from each side has further reduced the VSWR. Similarly, rows 1ˢᵗ, 2ⁿᵈ, 7ᵗℎ and 8ᵗʰ are terminated by 200 Ω resistors where all other elements are excited by 200 Ω feed lines. No improvements have been achieved on the VSWR of the active elements. Therefore, terminating the row does not improve the matching. Finally, columns 1ˢᵗ, 2ⁿᵈ, 3ʳᵈ, 6ᵗʰ, 7ᵗʰ, and 8ᵗʰ are terminated by 200 Ω resistors, where all other elements are fed by 200 Ω lines. The VSWR of the active elements has degraded especially at the lower frequency band as shown in Figure 3.32 (f).

In conclusion, it is demonstrated that resistive termination of the edge elements can significantly improve the VSWR of the array, yet, with respect to the linear dipole array in Figure 3.31(a), resistive termination of the edge rows has no significant effect on the VSWR value of the active elements. On the contrary, resistive termination of the edge columns lowers effectively the VSWR of the excited elements. In addition, the results in Figure 3.32 demonstrate that there is limit for the number of the edged columns, where aggressive termination beyond this limit will eventually degrade rather than improve the array bandwidth. This limit varies based on the array size, however, it is recommended that >40% should be active for broadband performance of the array.

Figure 3.32 shows that resistive termination of four edge columns (two from each side) of the array, yields significant improvements in the active VSWR, but, the high match
come at the expense of power dissipation in the resistors resulting in low total efficiency and thus, degrading the gains. To mitigate this problem, we examine both short and open edge termination. Similar to the resistive termination in case (d) of Figure 3.32, four edge columns, 1st, 2nd, 7th and 8th, will be terminated. The result from short-circuit, open-circuit and resistive edge termination will be compared with respect to total efficiency and corporate network VSWR.

**Short- / open-circuit terminations of edge elements:**

Figure 3.33 compares between the resistive, short-circuit and open-circuit terminations. As shown in (a) the corporate VSWR of the resistively terminated edges give the widest bandwidth. Followed by the short-circuit Edge-termination. However, as shown in (b), the resistive termination gives the widest bandwidth at the expense of total efficiency that is significantly lower than open-and short-circuited terminations.

![Comparison between resistive, short-circuit and open-circuit termination](image)

Figure 3.33: Comparison between resistive, short-circuit and open-circuit termination. (a) VSWR, and (b) total efficiency.
It is observed that in terms of array efficiency the short-circuited termination yields a much better performance when compared to the open circuit terminations. This more predominant at low frequencies where in the short circuit case the array efficiency is 78% compared with only 52% in open-circuit case. However, at high frequencies the open-circuit terminations seems to be more effective. The array performance when using combined short-and open-circuit terminations lays between the two cases as shown in Figure 3.34. It is noticed that the short-circuit termination give the optimum performance.

![Figure 3.34](image)

Figure 3.34: Comparison between short-circuit, open-circuit and short-/open-circuit termination. (a) VSWR, and (b) total efficiency.

The final 4×4 array with edge termination of two columns of shorted elements at each side is shown in Figure 3.35 (a). The VSWR of the edged array compared with the 4×4 with no edge terminations has improved.
Figure 3.35: (a) 4×4 array with short edge termination, and (b) VSWR comparison between non-terminated and terminated 4×4 array.
Next, the beam scan performance of the array is examined using the full wave simulator HFSS v.15.0. The PML boundary condition is used as shown in Figure 3.36. The array is fed through a lumped port, with 75Ω impedance, at each unit cell. Figure 3.37 presents the realized gain and directivity at broadside. Maximum gain of 10dB is achieved. The realized gain is lower than the directivity by less than 0.5dB. This is referred to the high match of the array.

Furthermore, the beam scan performance of the 4×4 array is examined. E-scan and H-scan are performed apart from each others. In each case, each unit cell is fed with a certain delayed signal to support beam steering at the desired direction. Figure 3.38 presents the beam scan performance in both E- and H-planes at three different frequencies (18, 28 and 40GHz). At each frequency, the beam is scanned with in the range ±45° with 15° step. As shown in Figure 3.38, H-plane scan proofs is symmetric, and the side lobe level is below -10dB. Whereas, E-plane scan has ripple due to finite nature of the ground plane. The ground plane radiates a field at certain frequencies that corresponds to the ground dimensions, this field interferes with the radiated pattern. Furthermore, to achieve a more clear beam scan performance it is desired to simulate larger number of elements.
Figure 3.36: Simulation setup of the 4×4 array prototype.

Figure 3.37: Simulated realized gain and directivity of the 4×4 array.
Figure 3.38: Beam-scan performance of the $4 \times 4$ array prototype.
3.7 Conclusion

In this section, versions of the coupled dipole array (CDA) with integrated feeds were presented. The challenges of feeding UWB phased array were reviewed, and the suitability of the unit feeding structure for brick-constructed array is examined. Marchand-balun is presented as a suitable feeding structure for wideband performance. The development of wideband microstrip line folded marchand balun for the CDA was discussed and the final design was presented. Finally, a proof of concept 4×4 prototype is simulated. The suitable edge termination type and number of terminated elements is determined. It was shown that the design is matched a cross ~2.3:1 bandwidth (VSWR<2.5, broadside) and exhibit good scanning capability ±45°.
Chapter 4: Feeding Network for Wideband Agile Beam Forming

It was shown in chapter I that there are different ways to deliver the desired current to each element for beam steering in phased arrays. Space-fed active lens, and reflect arrays [44] [45] are low profile, cheap way of implementing beam steering in phased arrays. This approach, however, suffers from instantaneous bandwidth that is limited by the use of phase control at the illuminated aperture. Therefore, for large arrays with modest bandwidth and wideband arrays, constrained feed is desired. Specifically, corporate feed, among the series one, which guarantees equiphase signal distribution for wideband arrays. Furthermore, it produces beams with frequency-independent scan angles.

On the implementation technology side, there are currently three technologies that exist in the implementation of the beam controlling circuit. Analog beam controlling technology is the most mature among digital and optical ones. It is a function of the well-established and it is still advancing fast with solid-state integration. Yet, the drawback of analog control is that it suffers of high loss. To reduce the loss, solid-state T/R modules are used at each element, or at sub array level to reduce the loss [42] combined with the appropriate RF beam steering architecture. Generally, in large arrays, it is optimum to have some sort of phase shifting at the unit cell level. After that, each sub array is backed by the desired TDU and controlling units. At the element level, TDU are used for wideband arrays, whereas, if the array is modest or small band, phase shifter can be
utilized instead of the TDUs. Compared to TDU, the phase shifter comes at a much lower profile.

In this work, we design corporate feeding network for beam controlling over modest bandwidth of a large phased array. The corporate network is implemented using analog technology. Phase shifter is implemented at each unit cell. Whereas, TDU can be used later at the sub array level. Going against the conventional approaches, in this design, we moved the phase shifter on top of the ground plane of the array to be integrated at the backside of the unit cell. This reduces the size and complexity of the feeding network. The effect of changing the position of the phase shifter on the noise figure of the design will be presented at the end.

In the following section, we propose the phase shifter design to be integrated at the backside of each unit cell of the phased array. The phase shifter is implemented using switched network topology, and low-pass/high-pass filter networks are used. Section 4.2, presents the integration of the phase shifter at the unit cell, and in the same section we examined the beam scan performance using the integrated phase shifter. Section 4.3 talks about the design procedures of the corporate feeding network. Full design performance including the feeding network and the antenna design is presented in Section 4.4. In section 4.5, we conclude this chapter.

### 4.1 Design of the Two-Bit MEMS Phase Shifter at Ka-Band.

Phase shifters at microwave and millimeter-wave are essential components in phased array antenna in radar, satellite and telecommunications. Phase shifters are currently based on ferrite materials, FET switched or BIN diodes. PIN diodes consume more DC
power than FET-based ones, however, it provide lower loss especially at high frequency. The ferrite based phase shifters perform well and could handle a lot of power, however, they desire manual tuning, consume a lot of DC power and are very expensive. On the other hand, solid-state phase shifters are compatible in shape with microwave components. They consume less DC power than ferrite and provide lower loss. Furthermore, they can be integrated with amplifiers on the same chip, and reduce the assembly cost of the phased array system [76] because they offer smart planar solution at microwave and millimeter-wave frequencies, they have been used largely in modern phased array systems.

4.1.1 Solid-State Phase Sifters

Analog and digital solid-state phase shifters are available. Analog solid-state phase shifters are achieved using Varactor diode that provides continuously variable phase shift. Digital phase shifters, with discrete set of phase delays, are implemented by using switches. The scan resolution and the side lobe levels of a phased array is related to the number of bits employed by the phase shifter, Most systems require a 3-bit or a 4-bit design. Modern solid-state phase shifters can provide 5 to 6 bits phase shifters for high performance systems at microwave frequencies. Both PIN diodes and FETs expand over 1–100GHz band, although, at this high frequency, the loss reaches up to 10dB. The switching time of solid state phase shifters is 1-50ns, depending on the RF power and size of the FETs and PIN diodes used. Solid-state phase shifters that utilize micro-electro mechanical technology in their design show lower loss and more compactness.
4.1.2 Design Topologies of Solid-State Phase Shifters

There are different topologies that are available in the design of a solid-state phase shifter, mainly; reflector type, switched line, loaded-line, Varactor and switched capacitor-bank, and phase shifters that are based on switched network, 1:N switches and antenna feeds [76].

Reflection type phase shifter is implemented using successive switches, series or parallel, on a transmission line [80]. Each switch, when locked, provides true time delay that equals double the phase delay per unit length. Reflection type phase shifter can be designed accurately up to 40GHz for 2-bit design. 3-dB coupler can be used with the reflect line resulting in a transmit type design. The bandwidth narrowed as the coupler see different impedance for every phase bit.

Switched line phase shifter is one of the simplest ways of implementing true time delay. It offers digital phase shifters using switched delay line technique. In this phase shifter, each bit is implemented separately and different phase delays are obtained by switching the required number of bits. Switched line phase shifter can provide very wide bandwidth (DC-50GHz), furthermore, it is simple to design and fabricate. The only drawback of this type is its large occupancy size.

Loaded line phase shifter was developed during 1960s -70s [116] [117] [118] [119]. The idea was to load a transmission lines with different impedances having the phase difference between the different loads to be controlled by the value of the loading impedance. Capacitive and inductive loads are used for phase delay and phase advance.
respectively. The bandwidth is relatively narrow <30%. This design is excellent for small delays.

Varactor and switched capacitor-bank phase shifter is standard analog design of the phase shifter at microwave and mmW frequencies. The design consists of 3-dB coupler and a reflection-type phase shifter. MEMS switched capacitor bank is used up to 6GHz and demonstrated 1.5-6:1 [120], however, this type does not work well above X-band.

Phase shifter based on switched networks is a common phase shifter design. It consists of two networks. The delay is achieved by passing the signal through either network. A commonly used network design in this type of phase shifter is the low-pass/high-pass filter network. The low-pass results in phase delays, whereas, the high-pass results in phase advance. The switched network design results in a much smaller design than the switched line type. Moreover, it has gained a popularity at the range 6-35GHz [121] [122] [123].

Phase shifter based on 1:N Switches consists of two switches, one at the input and another at the output to select between different delay lines. The delay line can be either LC network or transmission line. The advantage of this design is that the signal only passes two times through switches and not four times as in the standard 2-bits design; the main challenge being the input port match.

Finally, phase shifter based on antenna feeds is one in which it relies on switched balun feed and dipole antenna to achieve 0/180° phase shifter. Simple, when the signal flow is switched from right to left through the balun, the input voltage at the dipole feed point switches by 180°. However, this design results in a relatively narrow bandwidth (~10%).
therefore, it is not suitable for wideband antennas such as the spiral antenna or the tapered slot.

4.1.3 The Switched Network Topology

Switched network is a common phase shifter design at 6-35GHz [121] [122] [123]. The signal is derived into either network 1 or 2. Each has significantly low insertions loss and specific phase delay $\phi_1$ and $\phi_2$ respectively. A very common type of the switched network phase shifter is the low-pass/high-pass filter configuration shown in Figure 4.1. The high-pass filter gives phase advance, whereas, the low-pass filter creates phase delay. The capacitor and inductors in the circuit can be implemented through different ways; lumped components, transmission lines; or a combination of both. The high-pass/low-pass design results in a much smaller chip size than the switched line phase shifter.

![Figure 4.1: switched-filter phase shifter using (a) π-network, and (b) T-network.](image-url)
The design equations for π- or T-network using lumped LC components are described in Table 4.1. Both designs result in a maximum phase shift of 180°. Different phase shift values can be achieved through cascading multiple filter networks in series. The high-pass/low-pass filter design can be implemented easily using MEMS switches leading to a compact design [122] [121] [124] and the wideband all-pass switched network proposed in [125] [126].

Table 4.1: design equation of switched-filter phase shifter [127]

<table>
<thead>
<tr>
<th>π-network</th>
<th>T-network</th>
</tr>
</thead>
<tbody>
<tr>
<td>$X_n = \sin\left(\frac{\Delta \phi}{2}\right)$</td>
<td>$X_n = \tan\left(\frac{\Delta \phi}{4}\right)$</td>
</tr>
<tr>
<td>$B_n = \tan\left(\frac{\Delta \phi}{4}\right)$</td>
<td>$B_n = \sin\left(\frac{\Delta \phi}{2}\right)$</td>
</tr>
</tbody>
</table>

Low-pass filter:

$$L = \frac{X_n Z_o}{\omega}, \quad C = \frac{B_n}{\omega Z_o}$$

Low-pass filter:

$$C = \frac{1}{\omega X_n Z_o}, \quad L = \frac{Z_o}{\omega B_n}$$

**4.1.4 Two-Bits Tunable Switched Network MEMS Phase Shifter**

The phase shifter to be integrated with the presented phase array is provided by XCOM Co. It is designed to serve certain channels at the Ka-band, therefore, it does not operate over the entire bandwidth of the array. The phase shifter is implemented using high-pass/low-pass filters. The circuit model of the phase shifter is presented in Figure 4.2 (a). The phase shifter consists of two bits: an 180° bit and a 90° bit. The 180° bit is made of
T-network high-pass filter that gives 90° phase advance, and a π-network low-pass filter that gives 90° phase delay. Similarly, a 90° bit is made of two high-pass and low-pass networks with 45° phase advance and 45° phase delay respectively. The circuit model of

<table>
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<th>Value</th>
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</thead>
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</tr>
<tr>
<td>$C_2$</td>
<td>0.088 pF</td>
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<tr>
<td>$C_3$</td>
<td>0.214 pF</td>
</tr>
<tr>
<td>$C_4$</td>
<td>0.037 pF</td>
</tr>
<tr>
<td>$L_1$</td>
<td>0.5 nH</td>
</tr>
<tr>
<td>$L_2$</td>
<td>0.5 nH</td>
</tr>
<tr>
<td>$L_3$</td>
<td>0.7 nH</td>
</tr>
<tr>
<td>$L_4$</td>
<td>0.35 nH</td>
</tr>
</tbody>
</table>

Figure 4.2 (a) two-bits switched network phase shifter using high-pass/low-pass filter configurations, and (b) input reflection coefficient performance of the filter networks.
the phase shifter is designed using AWR software. The phase shifter proofs an impedance bandwidth of ~1.7:1 (18-30GHz) as shown in Figure 4.2 (b). The phase behaviors versus frequency for each filter design are shown in Figure 4.3. At 24GHz (mid of the band), the HPF of 180° bit gives 90° of phase shift (advance), whereas, the low-pass circuit in the same bit provide -90° of phase shift (delay) as shown in (a). Similarity, the 90° bit gives 45° phase advance when HPF network is activated, whereas, -45° phase (delay) when LPF network is activated. The performance of the 90° bit is shown in (b). The values of the LC lumped parameters are determined based on the equation in Table 4.1. The values are presented in the table of Figure 4.2 (a).

After the circuit model is achieved, the design is implemented using a full wave simulator. Tunable Piezo-electric capacitances are used with planar coil inductors. The diameter and number of turns of the planar inductor are determined based on [128]. For each filter network design, the simulation performance is curve fit with the circuit model to achieve the optimum value of the capacitors.
Figure 4.3: phase delay vs. frequency of the (a) $180^\circ$ and (b) $90^\circ$ high-pass/low-pass networks.
The two-bit tunable switched network MEMS phase shifter from XCOM Co. is shown in Figure 4.4. It consists of MEMS switches to select between different networks. The phase behavior versus frequency for the simulated filter networks is shown in Figure 4.5. The simulated results are compared with the results of the circuit model. The piezoelectric capacitances can be tuned to give continuous phase change during the specific range of each filter. For instance, the 180° LPF/HPF is designed to create any phase change within the range -90°/90°. This is simply done by tuning the piezoelectric capacitors in each filter network. MEMS tunable capacitors can be also used instead of the piezoelectric one. Notice, the control circuit of the phase shifter, including the bias lines and DC voltage courses are not discussed here nor included in the design.

Figure 4.4: 2-bit MEMS phase shifter, designed by XCOM Co., to be integrated with each array element.
Figure 4.5: Simulated performance of the 2-bit phase shifter by XCOM Co, compared with the circuit model.
4.2 Integration of the Phase Shifter with Simulated Array

Performance

The phase shifter is determined to be integrated at the backside of each unit cell. This novel approach is adapted to reduce the size and complexity of the design; furthermore, it reduces the overall loss of the feeding network. Figure 4.6 presents the new approach compared to the conventional method. As shown in (a) usually, the phase shifter is attached with each element after the LNA using cables and connectors. Whereas, in the proposed model shown in (b) we are moving the phase shifter at the unit cell directly. In this case, the LNA is moved at after the power divider network. Therefore, in the proposed model, there are no cables or transmission lines connecting the phase shifter to the previous or later devices. This should reduce the loss and cost of the feeding network. However, it is expected to increase the noise figure. The proposed model will be evaluated later in this chapter, and it will be compared with the conventional approach.

The proposed MEMS phase shifter is examined for beam forming. We prepared an array with integrated MEMS phase shifter at the back of each unit cell element. Due to complexity and memory limitations in simulations, a $4 \times \infty$-element model was considered for $45^\circ$ scan angle for both $E$- and $H$-planes as shown in Figure 4.7. At the backside of
Figure 4.6: RF beamformers (a) the Conventional, and (b) the proposed model.
each cell of the array, phase shifter with a certain switched network arrangement is integrated. For the first unit cell, we activated the track that goes through 90° LPF and 180° HPF networks. This track has the capability to give a maximum phase delay of -135°. Similarly, the second, third and fourth unit cells are integrated with phase shifter with 90°HPF\180°LPF, 90°LPF\180°HPF and 90°HPF\180°HPF respectively. The different tracks are tuned to beam scan equals 45° in E- and H-planes. The excited phase difference between the excessive elements is calculated according to equation:

$$\beta = \frac{2\pi}{\lambda} d \sin \theta$$

(4.1)

Where, $\beta$ is the phase excitation difference, $d$ is the element separation, $\lambda$ wavelength in free space and $\theta$ is the beam scan angle measured relative to broadside.

The beam scan performance is examined at three different frequencies, 18, 24 and 30GHz that covers the whole band of the phase shifter. As shown in Figure 4.7 the array beam can be scanned over the entire VSWR bandwidth (18-30GHz) with proper tuning of the MEMS phase shifters. The beam is slight shifted among the desired 45°. This is common in pahsed arrays with small number of elemetns. A better preformance is expected from an array of larger number of unit cells. In general, the beam is successfully steered with no degradation over the performance of the array, also, no beam squint is observed. Finally, the sidelobe level in H-scan shown in (b) and E-scan shown in (c) is better than -10dB.
Figure 4.7: (a) 4×Infinite TCDA with integrated MEMS phase shifters. (b) Simulated 45° beam scan performance in $H$-plane. (c) Simulated 45° beam scan performance in $E$-plane.
4.3 Design of the Corporate Power Division Network

Corporate network delivers the desired signal to each element that guarantees equiphase signal distribution for wideband arrays. For a 4×4 array, the corporate network consists of a 4:1 power divider. One of the concerns of designing this power divider at Ka-band is the loss. Material loss is high at Ka-band; therefore, one should make transmission lines short as possible, and to create them using low loss topology. Furthermore, there are different power divider approaches available. For phase array, adapting the approach that generates low reflection and negligible coupling between the ports is desired. High reflection alters the radiated field and might interfere with the original radiation, also, coupling will affect the radiated signal at each unit cell. A wide variety of waveguide power dividers were invented and characterized in the 1940s. In the mid-1950s-60s, many of these waveguides were redesigned to use stripline or microstrip line technology. One of the most common and simplest formats is the T-junction divider [129]. T-junction divider is lossless. It can be designed to provide any desired power ratio at the outputs. The phase at the ports is equal. However, T-junction divider cannot be matched at all ports. Furthermore, the coupling level between the different ports is relatively high. Another approach would be using the resistive divider [129]. Using resistance loads at the transmission lines of each port, the resistive divider offers matched ports, but still the isolation between the ports is poor. A lower loss alternative to the resistive divider is the Wilkinson divider [130]. Wilkinson divider is a common divider design. It offers matched ports and extremely low coupling (high isolation) between the ports. It adds relatively low loss compared to the resistive divider, however, it adds significant features.
to the feeding network of the phased array. Such as, high isolation between the ports and matched ports.

The Wilkinson design will be feeding the phased array. Therefore, a planar Wilkinson that is printed on Rogers3010 dielectric layer with 254μm is considered. For the type of transmission line, we selected microstrip line that match with the balun at each unit cell of the array. A novel planar Ka-band Wilkinson divider utilizes microstrip line is proposed in the following section

### 4.3.1 Design of 2:1 Wilkinson Divider at Ka-Band

A traditional circuit mode of 2:1 Wilkinson divider is shown in Figure 4.8 (a). It consists of a pair of quarter-wave length transmission lines “TL” and an isolation resistance “R”.

For Wilkinson with all ports having the same impedance “Z₀”, the characteristic impedance of the transmission line and the insulating resistance value are $Z_T = \sqrt{2}Z_0$, and $R = 2Z_0$ respectively. However, in our case, it is desired to transform the impedance as well. The impedance of each unit cell of the coupled dipole array is 75Ω, whereas, most coaxial feeding cables have an impedance of 50Ω. Therefore, it is desired to make impedance transformation through the Wilkinson divider as well. This is simply done by changing the impedance of the quarter-wave transmission lines, the new value $Z_T \approx \sqrt{2}Z_0=90\Omega$. While the insulating resistance value remains the same $R = 2Z_0=150\Omega$, where, $Z_0$ this time is the port impedance of the terminals (either P₂ or P₃). The performance of the 2:1 Wilkinson divider is shown in Figure 4.8 (b). A single stage Wilkinson divider is capable of around 3:1BW at VSWR<2.
Figure 4.8: (a) Schematic layout of the traditional 2:1 Wilkinson Divider and (b) its performance.

For the 4×4 array, 4:1 Wilkinson divider is desired. Therefore, the 2:1 Wilkinson has to be extended. Unfortunately, an N-way Wilkinson power divider can be planar only when
N=2, and cannot be planar for N≥3. However, it is always applicable to connect multiple 2:1 Wilkinson dividers. Therefore, multiway divider can be realized by interconnecting multiple 2:1 dividers, for instance 4:1 Wilkinson divider can be achieved by connection three of the 2:1 Wilkinson dividers. In the next section, design procedures of 4:1 Wilkinson by connecting two 2:1 Wilkinson dividers are presented. As indicated, the output ports should remain tight to each other, so there would be no phase difference across the resistor.

4.3.2 Planar 4:1 Wilkinson Divider for the CDA

A planar 4:1 BW Wilkinson divider for CDA array is presented in this section. The divider is to be printed on a Rogers3010 dielectric layer with 254μm thickness. The divider is implemented using microstrip transmission line that is seamlessly integrated with the marchand balun. An N:1 Wilkinson divider, with N≥3, is difficult to achieve using planar construction. However, three planar 2:1 Wilkinson dividers can be connected resulting in a planar 4:1 Wilkinson divider as shown in Figure 4.9. In practice, connecting the dividers with each other conveys change in their behavior. Usually, transmission lines with impedance $Z_o$ are used to link the dividers. For simplicity, the lines are made of equal length. Notice the transmission lines that are used to connect the dividers with other circuit components have significant effect on the input impedance when they have impedance that equals to the port impedance.

In Wilkinson divider, the resistor plays significant role in providing isolation and output port match. In ideal case, the resistor should have no phase different between its terminals to provide high isolation.
Figure 4.9: Schematic layout of a planar 4:1 Wilkinson divider achieved by three 2:1 Wilkinson dividers linked through a transmission line.

The reflection coefficient of a four way Wilkinson divider is given in [131]:

\[ S_{11}(f) = \frac{Z_A(f) - Z_0}{Z_A(f) + Z_0} \]  \hspace{1cm} (4.2)

Where,

\[ Z_A(f) = \frac{Z_1 Z_x(f) - jZ_1 \tan \theta_1}{2 j Z_x(f) \tan \theta_1 + Z_1} \]  \hspace{1cm} (4.3)

And,

\[ Z_x(f) = \frac{Z_2 \frac{Z_B(f) - jZ_2 \tan \theta_2}{j Z_B(f) \tan \theta_2 + Z_2}} \]  \hspace{1cm} (4.4)

Whereas;

\( \theta \): is the electrical length of the transmission line.
$Z_A$: is the input impedance seen at the input of the 4-way divider.

$Z_B$: is the input impedance seen at the input of the two-way divider.

$Z_X$: is the impedance seen after the linking transmission line for the two-way divider as shown in Figure 4.9.

Similarly, the reflection coefficient at the output port can be calculated based on the following equations, assuming the two Wilkinson dividers are connected to a $Z_o$ load each.

\[
S_{22}(f) = S_{33}(f) = \frac{1}{8 \tan \theta_3^2 - j\frac{\sqrt{2}}{8} \tan \theta_3 - 3} \tag{4.5}
\]

Whereas, the isolation is given by:

\[
S_{23}(f) = S_{33}(f) = \frac{-2 - j\frac{\sqrt{2}}{2} \tan \theta_3}{8 \tan \theta_3^2 - j\frac{\sqrt{2}}{8} \tan \theta_3 - 3} \tag{4.6}
\]

A 4:1 Wilkinson divider are determined based on design equations. Following design criteria were used, Reflection $S_{11}<-10\text{dB}$ over 18-40GHz, Transmission $S_{12}~6\text{dB}$, and coupling $S_{23}<-10\text{dB}$ over 18-40GHz. Later, design features were modified to fulfill the PCB fabrication capabilities. Namely, the trace width is limited to minimum value of 75$\mu$m that corresponds to a microstrip line impedance of $<75\Omega$. The performance is degraded slightly, however, it still meets our desired requirements. The performance of the final PCB-compatible planar 4:1 Wilkinson divider is shown in Figure 4.10. And the final circuit values are presented in Table 4.2. A proof of concept planar 4:1 Wilkinson divider is simulated using HFSS V15.0 as shown in Figure 4.11(a), and optimum performance is shown in (b). Reflection $S_{11}<-10\text{dB}$ is achieved over 18-40GHz, However, reflection $S_{22}, S_{33}$ slightly increased at the branch ports due to coupling.
Similar, the coupling $S_{23}$ level is below -10dB except at 33+GHz, where the coupling increases to -7dB. The insertion loss $S_{12}$ of the 4:1 divider is better than -9dB.

Figure 4.10: The performance of the 4:1 Wilkinson divider.

Table 4.2: Optimized circuit values of MB-CDA

<table>
<thead>
<tr>
<th>Equivalent Circuit</th>
<th>Figure 4.9</th>
</tr>
</thead>
<tbody>
<tr>
<td>$TL_1$</td>
<td>~1mm</td>
</tr>
<tr>
<td>$TL_2$</td>
<td>~1mm</td>
</tr>
<tr>
<td>$TL_3$</td>
<td>~1mm</td>
</tr>
<tr>
<td>$Z_1$</td>
<td>65 Ω</td>
</tr>
<tr>
<td>$Z_2$</td>
<td>50 Ω</td>
</tr>
<tr>
<td>$Z_3$</td>
<td>75 Ω</td>
</tr>
<tr>
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<td>50 Ω</td>
</tr>
<tr>
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<td>100Ω</td>
</tr>
<tr>
<td>$P1$</td>
<td>50 Ω</td>
</tr>
<tr>
<td>$P2$</td>
<td>75 Ω</td>
</tr>
</tbody>
</table>
Figure 4.11: (a) Full geometry of the Planar 4:1 Wilkinson Divider, and (b) Simulated performance.
4.4 Full Design Combining Array and Feed

The corporate feeding network is integrated with the $4\times4$ CDA as shown in Figure 4.12 (a). The 4:1 power divider is made such that it can be seamlessly attached with the CDA on the same Rogers 3010 board. The microstrip of the balun in each unit cell is attached with the power divider port using the same microstrip line to avoid any reflection at the interconnection between the array and the corporate network, the width of the microstrip line is made equal. Due to complexity and memory limitations, the MEMS phase shifter is eliminated from the $4\times4$ array. The design, including the array and corporate feed, are simulated using Ansoft HFSSv15.0. An impedance bandwidth spanning over 18-40GHz at VSWR<2 are obtained as shown in Figure 4.12 (b). Furthermore, the beam performance is examined in both $E$- and $H$-planes as shown in (c) and (d) respectively. Well behaved pattern and low side lobe level $<-10$dB is achieved in $H$-plane. Whereas, in $E$-plane, ripple appears due to finite structure of the ground plane and limited number of unit cells used in the design. As will be shown later, this issue is resolved after increasing the number of unit cells to $8\times8$ or larger. By comparing the maximum gain in (c) and (d) of Figure 4.12 with the one in Figure 3.38 (before adding the feeding network), we figure that gain drop is contributed by the added corporate feeding network.
Figure 4.12: (a) 4×4 CDA with integrated corporate feeding structure, (b) array bandwidth, (c) Pattern at different frequencies in E-plane, and (d) pattern at different frequencies in H-plane.
Finally, in order to examine the efficiency of our proposed beam steering architecture, we calculate the noise figure of the proposed model, in which the phase shifter is integrated at the unit cell, and compare it with the conventional method, where the phase shifter comes after the LNA. In the conventional architecture shown in Figure 4.13 (a), the antenna is followed by an LNA. The phase shifter comes after the LNA and connected through feeding cable that adds loss, and size. At the end comes the corporate feeding network. On the other hand, the proposed model integrates between the antenna, balun and the phase shifter. The phase shifter is this architecture utilizes the transmission lines of the balun, therefore, the feeding cable between the phase shifter and the former device is eliminated. This reduces the loss and size as well. However, in the proposed architecture, the LNA is moved to the end after the corporate network. This leads to an increase in the noise figure value. Table 4.3 compares the noise figure between the conventional and the proposed model. As shown, the noise figure of the proposed model is lower than the conventional one. Although the LNA is delayed in the chain of the proposed architecture, the integration of the phase shifter at the unit element, and the elimination and simplifications of the feeding cables not only reduce the complexity but it also improves the noise figure performance of the architecture.
Figure 4.13: RF beam Formers; (a) the conventional architecture, and (b) the proposed architecture.

Table 4.3: Noise figure values of the conventional and proposed RF beam forming architectures.

<table>
<thead>
<tr>
<th>Conventional Architecture</th>
<th>Proposed Architecture</th>
</tr>
</thead>
<tbody>
<tr>
<td>Antenna +Balun</td>
<td>Antenna+Balun+MEMS Phase Shifter</td>
</tr>
<tr>
<td>Feeding cable 1</td>
<td>Feeding Cable 1</td>
</tr>
<tr>
<td>LNA</td>
<td>Corporate Feeding network</td>
</tr>
<tr>
<td>Feeding cable 2</td>
<td>Feeding Cable 2</td>
</tr>
<tr>
<td>Phase Shifter</td>
<td>LNA</td>
</tr>
<tr>
<td>Feeding cable 2</td>
<td></td>
</tr>
<tr>
<td>Corporate Feeding Network</td>
<td></td>
</tr>
<tr>
<td>Total NF ≈6.5dB</td>
<td>Total NF≈5dB</td>
</tr>
</tbody>
</table>
4.5 Conclusion

In this chapter, version of RF beam forming architecture is adapted for beam forming of the large array with modest bandwidth. The architecture required phase shifter at each unit cell element of the corporate network. A two-bit MEMS solid-state phase shifter is developed, using switched low-pass/high-pass network design, and utilized for a lower loss and more compact phase shifter design. For power division, a planar 4:1 Wilkinson divider that provides isolated matched ports is considered. Moreover, a novel implementation approach of the corporate network is developed. In this approach, the phase shifter is moved on top of the ground plane compared to the conventional method where the phase shifter is underneath the ground plane. This advancement allows for lower profile, reduces loss and complexity and leads to an improved noise figure value of the chain.
Chapter 5: Fabrication Assembly and Measurements

Array cost is a considerable factor in many applications. There has been a need for optimum choices in collecting, fabricating and assembling array elements that are compatible with various methods of grouping elements. Fabrication and assembly become more and more challenging at higher frequencies. Several affordable fabrication techniques exist; PCB, thin film, and thick film technologies. Nowadays, PCB fabrication offers a wide range of dielectric substrates such as the Rogers family, with a wide range of relative permittivity and extremely low loss. On the other hand, thin and thick film offer slightly more advanced capabilities at the expense of a higher cost. Currently, PCB fabrication technology is limited to a minimum metal-trace width of 75μm, whereas thin film can reach as low as 25μm. In all of these technologies, adding vias, holes, and increasing the number of dielectric layers, increases the cost of fabrication significantly. As the frequency increases, the antenna size is reduced. This created further fabrication challenges. In this chapter, we present a 4×4 fabricated prototype of the CDA. We propose assembly methods that support strain relief, and allow more space for the connectors and feeding cable to be successfully integrated with the design. Finally, the fabricated prototype is measured in the anechoic chamber at The Ohio State University while the measured performance is presented in the last section.
5.1 Design Fabrication Using PCB Fabrication Technique

The fabricated array is similar to the optimum design that is presented in the last chapter, with the exception of the resistors of the Wilkinson divider that are eliminated from the fabricated prototype. This is simply because the measurements will be performed at Broadside, and it is believed that there exists a small phase change between the currents at each port of the Wilkinson. As a reminder, each of the four elements is combined through a divider that is printed on the same board, primarily to overcome space limitations. Figure 5.1 depicts images of the prototype. As seen, each row of elements and Wilkinson feed are printed on a single Rogers 3010 ($\varepsilon_r=10.2$, $\tan\delta=0.0013$ @10GHz) Printed Circuit Board (PCB). Thus, the 4x4 array consists of four total boards.

![Figure 5.1: A single brick of the fabricated array. (a) Back view showing 4-dipole unit cells, and (b) front view showing feeding network consisting of Marchand-baluns and a 4:1 divider.](image-url)
5.2 Assembly using Brick Construction Method

Boards are prepared for assembly using the brick construction method that allows for more space for the integration of the phase shifter above ground plane [43]. Assembly steps of the 4×4 array are shown in Figure 5.2.

As discussed in the previous section, four elements with a 4:1 divider are printed on a single brick. A ground plane is printed on Duroid 5880 material of 3.175mm thickness. Slots are made through the ground for the insertions of the array bricks as shown in Step

Figure 5.2: 4×4 array prototype at seven different stages in assembly.
1. The first brick of the array is inserted at the first to the back slot as shown in 2. The connector is attached to it as seen in Step 3. Currently, the minimum connector size is around 4.2mm, usually found in a miniSMP connector. It is noted that bricks in $H$-plane are only 2.1mm apart from each other. Therefore, connectors from different bricks overlap. This issue was not faced in $E$-plane simply because 4 cells, with 8.4mm width, are attached with each connector. To resolve this, the connector position has to alternate every other brick as shown in Steps 4 and 5. Another metal layer is added to the Duroid 5880 from the bottom, at which the connectors are soldered for more train relief. Finally, an additional Duroid 5880 layer encapsulates the connectors from the bottom as shown in 7. The connectors are also soldered with this layer as well. This sandwich structure creates rigid and fixed connectors.

5.3 Measurements

The assembled design is mounted on a metal platform that carries the 4:1 Wilkinson power divider as well, as shown in Figure 5.3. The power divider and the array are connected through high frequency coaxial cables with exact same lengths (phase difference <1ps) to guarantee the same phase at all elements. The measurements are performed in the compact range anechoic chamber at The Ohio State University. The path loss and feed cables effects are deducted from the measurements of the antenna under test (AUT) through a calibration antenna. As shown in Figure 5.4, the 4×4 array supports an impedance bandwidth of 18-40GHz at VSWR <2.

Figure 5.5 compares the measured and simulated broadside gain of the fabricated 4x4 array. To obtain simulated realized gain, the 4×4 array is simulated without the external
Wilkinson divider using full wave simulation of the array. Simulation is conducted using Ansoft HFSS v15.0. For the measurements, an 18-40GHz 4:1 Wilkinson power divider is connected to the array to deliver equal-phase signal to all elements. The losses of the divider (≈2dB Max.) are removed from the measured gain curve. A standard gain horn is used to perform the gain calibration as described in [132]. As shown, the measured patterns match the simulations, and the array has ~8dB of realized gain at 38GHz.

Figure 5.3: Fabricated design.

Figure 5.4: Measured reflection coefficient of prototype 4×4 array.
Figure 5.5: Measured broadside realized gain of prototype 4×4 array.

Figure 5.6 gives measured and simulated scan patterns in $E$- and $H$-planes. In these measurements, uniform excitation is used and all patterns are normalized. Patterns at 18-38GHz are depicted representing most of the array band. Excellent agreement between simulated and measured patterns is witnessed. The pattern at 28GHz has ripples at broadside direction. This is due to the small number of elements that are used; however, this can be mitigated by increasing the number of elements, or by eliminating the edge effect using field absorbers. These results validate the design of the CDA presented earlier. They also show a small size implementation of the design provides wide bandwidth.
Figure 5.6: Measured $E$- and $H$-plane pattern of $4\times4$ array for various frequencies.
5.4 Conclusion

In this section the Ka-band coupled dipole array (CDA) is empirically validated. Firstly, fabrication of a $4 \times 4$ prototype of the array is discussed, followed by brick construction assembly. Measured results are presented, showing apparent agreement with simulations. It is shown that the $4 \times 4$ CDA operates across 18-40GHz, and a well-behaved pattern is evident across most of the band.
Chapter 6: Conclusions and Future Work

6.1 Summary and Conclusions

The first chapter of this document discussed the ever-increasing need for electronic beam steering, wideband and low profile phased array for further SATCOM and 5G applications at Ka-band, and the large potential of tightly coupled arrays in this field. This work presents innovative CDA design with co-designed feeding networks that address contemporary challenges in the implementation of electronic beam scan arrays.

In Chapter 2, the evolution of the tightly coupled array was presented. This employs Wheeler’s current sheet, and Munk’s coupled array. Additionally, different architecture and topologies of the beam forming network are surveyed. This is essential in finding the optimum, lowest loss, feeding architecture for a beam former at Ka-band. The lowest loss design implies a good combination of beam steering architecture and device technology. (Such as corporate power division network with phase shifter at each element for beam forming, and combined with solid-state devices). The system would be simplified further and the loss would be reduced if the phase shifter is integrated at the back side of the unit cell element. This is inspired by the on-chip antenna concept, only with a much lower cost of realization.

In Chapter 3, a coupled dipole array (CDA) at Ka-band is illustrated. The design employs Marchand-balun feeding at the unit cell for wider bandwidth enhancement. This design
addresses the need for fabrication simplicity and low-loss at the Ka-band. As discussed, the antenna is made using a single dielectric layer, and feed gap is used combined with inter-element capacitance to increase the bandwidth. Marchand-balun is co-designed with the unit cell element to increase the bandwidth, while also providing balanced differential feeding of the dipole. The unit cell is modified and spaced for the phase shifter that was presented in the later chapter. The CDA design concept was validated empirically with a simulated 4×4 array.

In Chapter 4, we continued the discussion on the feeding network. We started by illustrating the design of the solid-state phase shifter. A two-bit tunable MEMS phase shifter by XCOM. Inc. is used. This phase shifter is implemented using switched network design with low-pass/high-pass filter networks. The phase shifter showed low loss (<2.5dB) over the band. The phase shifter was mainly designed for moderate bandwidth large phased arrays. After the phase shifter was designed, we examined simulated beam scan performance at a single scan angle. The presented beam scan performance of the 4×infinite array using the phase shifter proved the applicability of integrating the phase shifter at the back side of the unit cell, with no significant effect on performance. After that, we made a corporate feeding network using a four-way Wilkinson divider that is known for its high isolation between the ports and excellent match with the feed. A full design incorporating array and power division network is validated through a full wave simulation of a 4×4 array design. The design provides a bandwidth of 18-40GHz (VSWR<2) and a good scanning capability of $\theta=45^\circ$. Chapter 4 presents a fabricated 4×4 prototype, essentially a design assembly using brick construction and intensive
measurements. The measurements agree with the simulation, and the pattern behaved well in the band of interest (18-40GHz).

6.2 Future Work

While this work addresses several technical challenges in implementation of electronic beam steering array at K-band, specifically the tightly coupled dipole array, some areas in which it may be expanded are suggested. Modifying the presented design for dual-polarized operation is perhaps the highest priority, since many of the aimed applications demand this. Fabrication of a larger prototype to improve the radiation behavior might be the next highest priority. Other research topics include; improving the bandwidth and scanning capability of the presented design; realizing a simpler, lower-loss feeding network for small aerial applications; and scaling the design to a higher frequency.

The suggested directions for expansion of this work are:

1. **Dual-Polarized Implementation of the CDA**: Many aimed applications of the presented CDA require dual-polarization operation. Modification of the presented brick constructed array to support dual-polarization is not easy. However, the design can be changed into tile construction. Tile construction is also promising for scaling to higher frequencies.

2. **Fabrication of 8×8 Prototype of the Design**: In the presented 4×4 array, a dip has appeared in the realized gain versus frequency at around 25GHz. This dip has resolved by increasing the number of elements in the simulation to 8×8 as shown by Figure 6.1. Furthermore, an open stub balun is found to radiate at higher frequencies,
leading to a higher sidelobes level in the direction of the balun in $E$-plane and deteriorated pattern. This deterioration can be mitigated by using shorted stub instead of the $\lambda/4$ open stub, and alternating the direction of the balun every other row. Furthermore, similar behavior of the short stub can be achieved using a radial stub instead of using the via, this maintains the lower cost. Figure 6.2 presents an 8x8 array, with radial stub balun, that is constructed by using alternated rows for symmetrical pattern.

Figure 6.1: Simulated array with larger, 8×8, number of elements and broadside realized gain showing no dip at 25GHz.

3. **Increasing the Bandwidth of the CDA:** Chapter 3 presented a single layer coupled dipole array. In this design, the inter-element capacitance has not been utilized to its full capability. A wider bandwidth greater than 6:1 can be achieved by using overlapping inter-element capacitance. This requires using a higher number of dielectric layers. Furthermore, for corresponding wideband beam scanning performance, the phase shifter at each unit cell must be replaced with a TDU unit. This TDU has to be made at extremely low profile.
Figure 6.2: An 8x8 CDA constructed using alternated rows and radial stub for symmetrical pattern; and the corresponding normalized radiation pattern in $E$- and $H$-plane at three different frequencies.

4. **Simpler and Lower Loss Feeding Network:** As presented in chapter 4, a feeding network adds significant loss and weight to the design. It is desired to find an alternative approach. Digital architecture would be a good solution if the narrow bandwidth issues are resolved. Furthermore, in the current design, one might consider integrating the LAN at the backside of the unit cell as well, which would improve the noise figure while also reducing the loss and weight of the feeding network.
5. **Scaling the Array to Fill the Gap Between Ka-band and W-band:** As of today, coupled dipole array has been realized at Ka-band. There is tremendous potential of the on-chip antenna at W-band and higher. On chip antenna technology is still relatively unfounded and extremely costly. It would be helpful to implement tightly coupled array using affordable fabrication techniques over the band Ka-W. Our proposed design in Chapter 3 indicates that the dimensions of the unit cell at Ka-band are close to the maximum capabilities of the PCB technology. Therefore, scaling the tightly coupled dipole array to a higher frequency will lead to a smaller design that is difficult to fabricate using PCB technology (75μm minimum trace width). Another solution would be switching to thick film technology that can drop to as low as 25μm minimum trace width.

6. **Wide-band true time delay that can be integrated at the unit cell:** an ultra-wide band performance of a phased array is irrelevant without developing a corresponding wideband phase shifter for beam steering. Right now, there are time delay units (TDUs) that are wideband. These delay lines are bulky. For compact design that can be integrated at the unit cell of the ultra-wideband tightly coupled array, it is desired to develop Lower-profile TDUs.
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