## Title

# Design of a Class of Antennas Utilizing MEMS, EBG and Septum Polarizers including Nearfield Coupling Analysis 

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# Design of a Class of Antennas Utilizing MEMS, EBG and Septum Polarizers including Near-field Coupling Analysis 

A dissertation submitted in partial satisfaction of the requirements for the degree Doctor of Philosophy in Electrical Engineering by

Ilkyu Kim

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Ilkyu Kim

# ABSTRACT OF THE DISSERTATION 

# Design of a Class of Antennas Utilizing MEMS, EBG and Septum Polarizers including Near-field Coupling Analysis 

by

Ilkyu Kim<br>Doctor of Philosophy in Electrical Engineering<br>University of California, Los Angeles, 2012<br>Professor Yahya Rahmat-Samii, Chair

Recent developments in mobile communications have led to an increased appearance of short-range communications and high data-rate signal transmission. New technologies provides the need for an accurate near-field coupling analysis and novel antenna designs. An ability to effectively estimate the coupling within the near-field region is required to realize short-range communications. Currently, two common techniques that are applicable to the near-field coupling problem are 1) integral form of coupling formula and 2) generalized Friis formula. These formulas are investigated with an emphasis on straightforward calculation and accuracy for various distances between the two antennas. The coupling formulas are computed for a variety of antennas, and several antenna configurations are evaluated through full-wave simulation and indoor measurement in order to validate these techniques. In addition, this research aims to design multifunctional and high performance antennas based on MEMS (Microelectromechanical

Systems) switches, EBG (Electromagnetic Bandgap) structures, and septum polarizers. A MEMS switch is incorporated into a slot loaded patch antenna to attain frequency reconfigurability. The resonant frequency of the patch antenna can be shifted using the MEM switch, which is actuated by the integrated bias networks. Furthermore, a high gain base-station antenna utilizing beam-tilting is designed to maximize gain for tilted beam applications. To realize this base-station antenna, an array of four dipole-EBG elements is constructed to implement a fixed down-tilt main beam with application in base station arrays. An improvement of the operating range with the EBG-dipole array is evaluated using a simple linkbudget analysis. The septum polarizer has been widely used in circularly polarized antenna systems due to its simple and compact design and high quality of circularity. In this research, the sigmoid function is used to smoothen the edge in the septum design, which makes it suitable for HPM systems. The PSO (Particle Swarm Optimization) technique is applied to the septum design to achieve a high performance antenna design. The electric field intensity above the septum is evaluated through the simulation and its properties are compared to simple half-plane scattering phenomena.

The dissertation of Ilkyu Kim is approved.

| Yuanxun Ethan Wang |
| ---: |
| Christoph Niemann |
| Yahya Rahmat-Samii, Committee Chair Candler |

To my lovely family
who provides unconditional support and constant encouragement.

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I. Kim and Y. Rahmat-Samii, "Revisiting stepped septum circular polarizer using fullWave simulations," 2011 IEEE Antennas and Propagation International Symposium, 2011.
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## Chapter 1

## Introduction

With the growing demand for modern communication systems, multiple functionalities and high performance antenna systems become more desirable. The reconfigurable antenna allows the operational flexibility incorporated into the antenna such as reconfigurable resonant frequencies or adaptive radiation patterns. For instance, mobile antennas incorporated with RF switches can change its resonant frequency or base-station antenna utilizes the down-tilt main beam toward a mobile. In addition to the multifunctional antennas, the circularly polarized antennas with high power handling are required to realize some of satellite communications and military applications. These functionalities can be incorporated into the feed horn design for a large reflector antenna which transmits high power signals. In parallel to the novel antenna designs, many of current communication systems require an ability to estimate the linkbudget between multiple antennas. Efficient communication system design can be achieved through an accurate linkbudget evaluation that ranges from the near-field to the far-field. This research is aimed at investigating the novel antenna design, based on MEMS (Microelectromechanical Systems) and EBG (Elecromagnetic Bandgap), and the circular
polarizer to realize a high performance horn antenna, and linkbudget evaluation which can be applicable to the wide ranges of distances and antennas. An overview of all the researches is depicted in Figure 1.1


Research areas in this dissertation


Figure 1.1: A schematic diagram of the areas covered in the dissertation.

## 1. 1 Linkbudget Evaluation in the Near-field

It is crucial to determine the accurate operating ranges of those devices in new areas, medical wireless imaging and monitoring [1]. The strict requirements on the electronic ticket payment system [2] and a wireless power transfer [3-7] have been emphasized with respect to the interference and the efficiency in the system. With the increasing demand for the wireless communications, more antennas are designed to work at close distances. The capability to estimate the power transmission between antennas in the near-field becomes more important. A simple and effective formula that enables accuracy prediction in the near-field is most required.

Several research activities have been reported to accurately predict the power transmission between closely spaced antennas, in particular, within a radiating near-field region. Common techniques for estimating the coupling can be categorized in three ways:

1) Friis formula including a correction term.
2) Integral form of the coupling formula.
3) Numerical evaluation using a full-wave simulation.

Figure.1.2 depicts the flowchart for estimating near-field power transmission. This dissertation presents an enhancement in the Friis and coupling formula, and coupling evaluation using full-wave simulations and indoor measurement. Those methods can be used depending on 1) flexibility in antenna configuration, 2) required information, 3) computational complexity. The generalized Friis formula [8-9] is advantageous due to its simple form of formula and simple required information such as gain at bore-sight and
operational frequency, but its effectiveness is limited to the Fresnel region. On the other hand, integral form of coupling formula [10-11] computes the integral of the products between two far-field patterns, which can be applied to the entire radiating near-field region with moderate computation complexity. Lastly, the numerical evaluation is most accurate method although it requires the highest computational complexity. The features for each method are summarized in Figure 1.3, and will be discussed in this dissertation.


Figure 1.2: Flowchart for the linkbudget evaluation in the near-field communication:

1) generalized Friis formula, 2) integral coupling formula, and 3) numerical evaluation.

|  | Computational <br> complexity | Required <br> information | Flexibility in <br> configuration |
| :---: | :---: | :---: | :---: |
| Generalized <br> Friis formula | Low | Low | Low |
| Integral coupling <br> formula | Low | High | High |
| Numerical <br> evaluation | High | Low | High |

Figure 1.3: Advantages and disadvantages of the three techniques.

### 1.2 MEMS Reconfigurable Antenna

In recent past, RF switches are incorporated into antenna configurations to attain more diverse functionalities [12-17]. A two-terminal diode switch has been effectively used due to its wide availability [18]. Although the implementation of the diode switch in the antenna design is relatively simple, diode switches usually suffer from relatively high insertion loss and low linearity [19]. With the growing interest in the high frequency application, the RF operation is strongly influenced by the switch characteristic. In this aspect, RF MEMS switch would be a better candidate for designing reconfigurable devices because of relatively low loss and high isolation features [20-23]. The characteristics of the MEMS switch and setup of the measurement are shown in Figure 1.4. The MEMS switch shows $0.2-0.5 \mathrm{~dB}$ insertion loss and $20-30 \mathrm{~dB}$ isolation level across $4-6 \mathrm{GHz}$. This feature is advantageous to design the MEMS reconfigurable antenna. One disadvantage of the MEMS switch is the relatively high operational voltage required, but recent research activities on MEMS reconfigurable antennas have reported antennas operated within reasonable voltage ranges. One challenge in this MEMS switch
is an inclusion of a three-terminal commercial packaged MEMS switch into the antenna in designing reconfigurable antennas. In particular, when MEMS switches are incorporated into complicated antenna structure, the design of the bias network becomes more difficult. To overcome this problem, an enhancement to the bias network design is required. In this dissertation, the bias network suited for slot loaded patch antenna will be presented in Chapter 3.


Figure 1.4: Setup of measurement for MEMS switch and characteristic of the MEMS switch.

### 1.3 Beam-tilting EBG-dipole Array Base-station Antenna

Beam-titling base-station applications have been widely used in order to realize an efficient wireless communication networks. With the radiation pattern that can be tilted towards a desired direction, the system capacity of the base station like applications can be greatly improved [24-26]. The fixed down-tilt of the main beam is usually employed in the common base station, thus the receive signal toward a mobile device can be increased. Beam-tilting base stations can be achieved with the array antennas where phase to the the array antenna is electronically controlled [27-29] or the position of the array is mechanically adjusted [30-32] as shown in Figure 1.5. For the electronically controlled arrays, the varactor and RF MEMS actuated phase shifter has been implemented in the array design to provide a flexible beam-tilting ability. This array antenna, however, has some limitations to achieve a high directivity across a wide angular range. Specifically, one disadvantage in this array is non-changed element pattern while steering the array pattern with a progressive phase. It usually leads to a huge gain loss in the total radiation pattern when the beam is steered toward a desired direction. Another alternative is the beam-tilting base station with a mechanical movement. This technique would be more desirable due to no gain loss in the radiation pattern. This is because the maximum array pattern matches to the element pattern. In spite of the advantage in this array antenna, there still remains a problem in the mechanical installation resulting from positioning the inclined array. To overcome problems addressed and to satisfy the strict base-station requirements, advanced base-station design is required to achieve a future base station array antenna. To achieve high performance array antenna, incorporating beam-tilting single element into array antenna might be a
solution.

EBG (Electromagnetic Band Gap) structure [33-37] as shown in Figure 1.6 can be a good candidate to realize the single element. This is because EBG structure is suitable for a low-profile wire antenna that can form the broadside radiation pattern with a relatively high directivity. In particular, it can change the radiation pattern towards a tilted direction by modifying the EBG unit cells. In this dissertation, the design of EBG-dipole array antenna will be presented with the base-station like antennas performance .


Figure 1.5: Examples of mechanically or electronically controlled array antenna.


Standard EBG structure
Figure 1.6: Configuration of the EBG structure.

## 1. 4 Smooth Sigmoid Profiled Circular Polarizer: Septum

Another research area in this dissertation is the circular polarizer, septum. A CP antenna system has been widely used due to its robustness against the fading and misalignment from which a linearly polarized antenna suffers [38-40]. To realize the reflector feed systems, it is also critical to produce the good circularity with compact size. The septum has gained a widespread popularity due to its simplicity, small size, and its high quality of circularity provided [41-49]. As shown in Figure 1, the septum is a simple three-port device placed within a horn antenna which can convert a linearly polarized excitation into a circular polarized wave. The waveguide may then feed a horn antenna to radiate CP waves for a reflector system or even a reflectarray. Another feature of the septum is that both polarizations (RHCP/LHCP) can be generated by a proper excitation of the ports. Depending on the system requirements, this can be used to create a circular polarization duplex system, where one polarization (e.g. RHCP) represents the transmitted signals and the other represents the received signals, as shown in Figure 1.7. This feature also allows the septum to operate as an Orthogonal Mode Transducer (OMT) to decompose a received wave into its respective RHCP/LHCP components or even its linear polarization components with some further processing after the septum. There have been many recent advances towards developing compact and practical high power sources, and implementation of these may be imminent for military or space applications such as communication systems, radar, or even non-lethal weapons systems [51-54].


Figure 1.7: Circular polarization and high power ground station reflector and proposed septum polarizer.

However, there remains limited RF devices that can operate at these power levels. Regardless of the advantages and simplicity of a septum, one inherent potential limitation in the standard septum design is the use of sharp corners. This can intensify the electric fields near the septum edges, ultimately leading to possible breakdown for high power applications. To realize a high power microwave (HPM) system using a septum, high power handling should be taken into consideration in its design. Handling HPM has been emphasized in the past for feed horn antennas in reflector antenna systems. Some effort
has been made in order to avoid both air breakdown and multipaction within horn antennas [54-55]. However, these works have been limited to linearly polarized antenna systems. Some relatively low power examples of deep-space radiometry and satellite communications using circular polarization can be found [56-57]. With practical high power microwave sources coming closer to realization, it will be important to provide RF devices that can handle high power, and therefore this dissertation will focus on achieving better power handling in the septum design.

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## Chapter 2

## Linkbudget Analysis in Near-field

In this chapter, the analytical formulas are presented to estimate an accurate coupling between antennas in near distance. One of the formulas presented is based on the asymptotic series form of the Friis formula, and the other one uses the integral form of the inner product of the two vector far-field pattern. Given minimum information such as far-field pattern, the separation distance and operating frequency, an accurate coupling between two antennas can be quickly determined with using these formulas. These formulas can be effectively applied to the very broad range of the antennas. The applications include the small dipole, standard waveguide, horn antenna, and large reflector antenna.

## 2. 1 Introduction

As discussed earlier, it is critical to accurately predict the coupling between two antennas placed within the near-field region. In the near-field region, the radiation pattern becomes broadened, and thus the gain of the antenna is reduced. This can be gauged by using gain reduction factor. Due to this reduction, the coupling between two antennas is always smaller than the Friis formula's estimation. Therefore, the accuracy of the formula can be
judged by how the improvement to the Friis formula is well estimated. To estimate the linkbudget in near-field region, three representative methods can be utilized: 1) Friis formula with correction term, 2) Integral form of coupling formula. In this chapter, the results from the proposed formulas are compared with the full-wave simulation and measured results so as to evaluate the formulas. Note that the power transmission between two antennas within a radiating near-field region is sometimes written as the terminology "coupling" in this chapter. The power transmission of diverse antennas is evaluated using the proposed formulas, full wave simulation, and measurement.

## 2. 2 Friis Formula with Correction Term

Asymptotic Friis formula is presented through analytical derivation. This intuitive asymptotic Friis formula makes it possible quickly determine the gain reduction for all existing antenna, in other words, universal gain reduction factor. With the proposed formula, the power transmission in a near-distance, especially in a Fresnel region can be effectively determined.

## 2. 2. 1 Theory of the Friis Formula with Correction Term

The Friis power transmission formula [1] has been widely used to calculate the link budget in wireless communications due to its effectiveness in practice and simplicity in calculation. Although it was originally derived using plane waves, Hogg [2] summarized four different methods of deriving the Friis formula, and all led to the same conclusion. The assumptions used in its derivations suggest that the formula is only valid in the far-


Figure 2.1: TX and RX antennas that are separated by distance R are depicted.


Figure 2.2: Power transmissions between two dipole antennas are obtained using Friis formula and full-wave simulation.
field region. Some efforts have been made to present appropriate models to estimate the power transmission levels between antennas in the Fresnel region [3-7]. Specifically, the concept of the gain reduction (or correction) factor [3-6, 8-10] was widely employed to improve the accuracy of the antenna gains used in the Friis formula. However, the applications of those proposed formulas are limited where most of them are focused on simple analytical antenna models, due to the lacking of numerical capabilities


Figure 2.3: On-axis power density of a uniform circular aperture with diameter of $10 \lambda$ using the exact analytical formula (Near Field), the formula with Fresnel approximation (Fresnel Field), and the formula with far-field approximation (Far Field).
at that time, and the scenario at boresight is only considered. Recently some research activities on the Friis formula were reported, but their effectiveness was limited to the farfield region [11-12].

A receiving antenna is located within the far-field or Fresnel region of transmitting antennas as shown in Figure 2.1. It is well known that the power transmission level $S_{21}$ (or $P_{r} / P_{t}$ ) between two antennas in the far-field region can be predicted by using the Friis formula. Its complete form is given by

$$
\begin{align*}
S_{21}\left(R, \theta_{r}, \phi_{r}\right)= & \frac{P_{r}}{P_{t}}=\frac{\lambda^{2}}{16 \pi^{2} R^{2}} G_{t}\left(\theta_{t}, \phi_{t}\right) G_{r}\left(\theta_{r}, \phi_{r}\right)  \tag{2.1}\\
& \quad \times\left(1-\left|\Gamma_{t}\right|^{2}\right)\left(1-\left|\Gamma_{r}\right|^{2}\right)\left|\hat{\rho}_{t} \cdot \hat{\rho}_{r}\right|^{2}
\end{align*}
$$

where $G(\theta, \phi)$ is the far-field gain pattern with $(\theta, \phi)$ angular dependence, $1-|\Gamma|^{2}$ is the input
impedance mismatch, and $\left|\hat{\rho}_{t} \cdot \hat{\rho}_{r}\right|^{2}$ is the polarization mismatch. Subscripts $t$ and $r$ refer to the transmitting (Tx) antenna and the receiving ( Rx ) antenna, respectively. $R$ is the separation distance between the phase centers of the $T x$ and $R x$ antennas that are placed at arbitrary position. Before discussing rotational scenarios, we focus only on the boresight scenario, and accordingly, the angular-dependent antenna gain pattern used in (1) is replaced by its far-field gain in the boresight direction.

Although the standard Friis formula is successful in the far-field region, the accuracy deteriorates considerably at near distances. Figure 2.2 shows a comparison between standard Friis formula and simulated power transmission of dipole antennas, and simulated result exhibits a huge deviation from the Friis formula which is equivalent to 5 dB at $R=0.3 \lambda$. The reason is that the formula only uses the far-field antenna gain, which may vary considerably in the Fresnel region at different distances from the antenna phase center. The key to improving its accuracy is to incorporate the antenna gain variations in the Fresnel region. By replacing the far-field antenna gains used in (2.1) with the antenna gains in the Fresnel region $G_{F}(R)$, a modified Friis formula is obtained,

$$
\begin{equation*}
S_{21}(R)=\frac{P_{r}}{P_{t}}=\frac{\lambda^{2}}{16 \pi^{2} R^{2}} G_{F, t}(R) G_{F, r}(R) \tag{2.2}
\end{equation*}
$$

Note that the antenna gain is now a function of $R$. Although in the Fresnel region the phase center slightly shifts for the gain patterns at different distances, $R$ is still measured from the phase center of the antenna's far-field gain pattern. A gain reduction factor $\gamma$ is usually employed to represent the gain decrease effect in the Fresnel region. It is usually defined as

$$
\begin{equation*}
\gamma(R)=\frac{G_{F}(R)}{G} \tag{2.3}
\end{equation*}
$$

the ratio of the antenna gains in the Fresnel and far-field regions. Thus (2.3) can be rewritten as

$$
\begin{equation*}
S_{21}(R)=\frac{P_{r}}{P_{t}}=\frac{\lambda^{2}}{16 \pi^{2} R^{2}} \gamma_{t}(R) G_{t} \gamma_{r}(R) G_{r} \tag{2.4}
\end{equation*}
$$

The assumption used in (2.3) and (2.4) is that the gain reduction factor of any individual antenna can be achieved through analytical derivation, numerical simulations or measurements of the antenna gain pattern in the Fresnel region. The antenna gain reduction factor may vary differently for different types of antennas; however, $[8,9,5]$ suggested an approximate quadratic variation of the antenna gain in the Fresnel region with respect to its far-field gain:

$$
\begin{equation*}
\gamma(\Delta)=\frac{G_{F}(\Delta)}{G}=1-\alpha \Delta^{-2}+O\left(\Delta^{-4}\right) \tag{2.5}
\end{equation*}
$$

where $\Delta$ is the normalized form of the separation distance $R$ from the phase center of the antenna's far-field gain pattern, and $\alpha$ is a unique coefficient that approximately determines the variation of the antenna gain at different distances in the Fresnel region.

In general, $\alpha$ is different for different antennas. It can be determined analytically or numerically [8, 9, 7]. It is crucial to observe that $\alpha$ also varies with different ways of normalizing the separation distance $R$. With a proper definition of the normalized separation distance $\Delta$, a universal coefficient for a broader range of antennas may be achieved. In this paper, we follow the quadratic asymptotic form of the gain reduction factor used in literature and propose a novel normalization form of separation distance $R$

$$
\begin{equation*}
\gamma(\Delta)=\frac{G_{F}(\Delta)}{G}=1-\alpha \Delta^{-2}, \Delta=\frac{R}{2 \lambda G / \pi^{2}} \tag{2.6}
\end{equation*}
$$

where $G$ is the antenna gain in the far-field region and $R$ is the separation distance. The
separation distance is normalized with respect to the antenna gain (or the effective area) and wavelength so that the effects of operating frequency, antenna size and tapering efficiency can all be included in $\Delta$, resulting in a more universal $\alpha$. For a uniform circular aperture field of diameter $D, \Delta=1$ also refers to its far-field edge defined by $2 D^{2} / \lambda$. The coefficient $\alpha$ can be determined by solving (2.6) where the antenna gains in the Fresnel and far-field regions are assumedly available through simulations or measurements. It is worth pointing out that (2.6) is valid only within the Fresnel region where the on-axis power density of the antenna decreases monotonically.

Figure 2.3 shows an example of normalized on-axis power densities of a uniform circular aperture with diameter of $10 \lambda$. Assuming the Fresnel approximation is utilized, it varies as [20]

$$
\begin{equation*}
|E|^{2}(R) \propto\left[1-\cos \left(\frac{2 D^{2} / \lambda}{8 R} \pi\right)\right] / 2 \tag{2.7}
\end{equation*}
$$

Therefore, the distance $R$ of the last peak value can be calculated by

$$
\begin{equation*}
R=\frac{1}{8} \frac{2 D^{2}}{\lambda}=\frac{A}{\lambda \pi} \tag{2.8}
\end{equation*}
$$

where $A$ is the physical area of the aperture. It is conjectured that $A$ in (2.8) can be replaced by its effective area $A^{e}$ in order to include the effects of tapering efficiency and phase errors which reduce the total aperture efficiency from unity. Supposing that no input impedance mismatch exists and the two antennas are aligned in the beam maximum direction with matched polarization, by using the definition of the antenna effective area

$$
\begin{equation*}
A^{e}=\frac{\lambda^{2} G}{4 \pi} \tag{2.9}
\end{equation*}
$$

(2.10) can be rewritten as

$$
\begin{equation*}
R=\frac{A^{e}}{\lambda \pi}=\frac{\lambda G}{4 \pi^{2}} \tag{2.10}
\end{equation*}
$$

Although (2.10) is not achieved through rigorous derivations, it does provide certain insightful guideline for the quick determination of the Fresnel region. The on-axis power densities for several analytical aperture fields can be found in [9] and they all confirm the validity of (2.6).

## 2. 2. 2 Universal Reduction Coefficient $\alpha$

An extensive study of several representative aperture antennas such as dipole, horn antennas and reflector antennas has been performed using FEKO simulations. FEKO is a powerful full-wave simulation tool suitable for analysis of the power transmission. The complexity of FEKO analysis is primarily determined by the antennas' electrical sizes, but almost not affected by the separation distance between them. The boresight power transmission levels between two identical antennas at various separation distances are obtained. It is assumed that antennas are both impedance matched and polarization matched. More than twelve different types of antennas are evaluated. Among them five examples are selected and required information such as the aperture dimension, far-field antenna gain, and the location of phase center are listed as below.

1) $K u$-band standard gain horn: Aperture size $=2.15 \lambda \times 1.59 \lambda(D=2.67 \lambda$ is the diameter of the circle enclosing the horn aperture) at 12.7 GHz , simulated far-field gain $=15.2 \mathrm{~dB}$, and phase center $=5 \mathrm{~mm}$ below the aperture.
2) C-band reflector antenna: $D=34.84 \lambda$ at 5.5 GHz , simulated far-field gain $=30.0 \mathrm{~dB}$, and phase center $=$ the center of the reflector rim.
3) 1.4 GHz half-wavelength dipole antenna: $D=\lambda / 2$ at 1.4 GHz , and simulated far-field gain $=2.14 \mathrm{~dB}$.
4) C-band circular horn: $D=3.0 \lambda$ at 5.5 GHz , simulated gain $=18.5 \mathrm{~dB}$, and phase center $=7 \mathrm{~mm}$ below the aperture.
5) C-band rectangular open-ended waveguide: $D=0.36 \lambda$ at 5.5 GHz , and simulated gain $=6.7 \mathrm{~dB}$.


Figure 2.4: Gain reduction factors of the representative antennas as a function of $R /\left(2 D^{2} / \lambda\right)$.


Figure 2.5: Gain reduction factors of the representative antennas as a function of

$$
R /\left(2 \lambda G / \pi^{2}\right)
$$



Figure 2.6: Gain reduction factors of the representative antennas as a function of

$$
R /\left(2 \lambda G_{\mathrm{FS}} / \pi^{2}\right)
$$



Figure 2.7: Gain reduction factors of the representative antennas as a function of

$$
R /\left(2 \lambda G_{\mathrm{A}} / \pi^{2}\right) .
$$

The gain reduction factors are calculated using the full-wave simulation results by invoking (2.4) and (2.6). In literature [4-5], the conventional normalization of $R$ was based on the far-field edge distance, $\left(2 D^{2} / \lambda\right)$. Both methods are used to obtain the gain reduction factors as a function of different $\Delta$ :

$$
\begin{align*}
\Delta & =\frac{R}{2 \lambda G / \pi^{2}}(\text { Proposed normalization })  \tag{2.11}\\
\Delta & =\frac{R}{2 D^{2} / \lambda}(\text { Conventional normalization }) \tag{2.12}
\end{align*}
$$

which are presented in Figure 2.4 and 2.5. Note that the simulated nearest distance for dipole antenna is different from other antennas due to the severe interaction between dipole antennas within a very close distance. It is observed that the proposed
normalization in (2.11) achieves more converged curves for aperture antennas such as horns and reflectors. This attributes to the effective area rather than physical area of the aperture used in the assumption to derive (2.11). However, in spite of that, the normalization shows the limitation for low-gain antennas such as small dipoles and openended waveguides.

The far-field gain $G$ used in (2.11) is based on the ideal aperture antenna that radiates toward one-side of the aperture. Therefore, it is reasonable for utilizing the one-side total radiated power when calculating directivity of the antenna. The modified gain, namely the front-side gain $G_{F S}$, can replace the far-field gain $G$ presented in (2.11). The front-side gain due to the front-side radiate power and associated gain reduction factor can be defined as

$$
\begin{equation*}
G_{F S}=\frac{4 \pi U_{m}}{P_{F}}, \quad \gamma_{F S}(\Delta)=1-\alpha_{U}\left(\frac{R}{2 \lambda G_{F S} / \pi^{2}}\right)^{-2} \tag{2.13}
\end{equation*}
$$

where $U_{m}$ is the maximum radiation intensity and $P_{F}$ is the total radiated power over a hemisphere where maximum radiation occurs. For low-gain antennas, there is a significant difference of total radiated power between half- and entire- sphere while similar total radiated power can be obtained for high-gain antennas. For example, in the case of Ku-band horn, modified gain $G_{F S}$ is obtained as 15.1 dB through calculating $P_{F}$ from the simulated pattern, which is similar to simulated far-field gain $G$ of 15.2 dB . For a dipole antenna, $G_{F S}$ is equivalent to $2 G$, twice of the far-field gain.

Figure 2.6 shows the gain reduction factors plotted using (2.13), and it is obvious that more converged gain reduction factors are obtained for both high- and low-gain antennas. By matching all the curves, an empirical universal coefficient $\alpha$ can be found to be $\alpha_{U}=$ 0.06 .

The full knowledge of an antenna's 3D radiation pattern may not be readily available to researchers under most circumstances. Therefore, a further simplification can be adopted to avoid the calculation of the front-side radiated power. The antenna gain is adjusted simply according to its category: low-gain or high-gain antennas as follows:

$$
\begin{align*}
& G_{A}=G(G \geq 10 d B: \operatorname{high} G)  \tag{2.14}\\
& \quad=G+3 d B(G \leq 10 d B: \operatorname{low} G) \quad, \gamma_{A}(\Delta)=1-\alpha_{U}\left(\frac{R}{2 \lambda G_{A} / \pi^{2}}\right)^{-2}
\end{align*}
$$

where $G_{A}$ is named the adjusted gain. Note that the adjusted gain $G_{A}$ is suggested after an extensive case study and no rigorous analytical derivation is currently available. As shown in Figure 2.7, the proposed method in (2.14) provides similar performance as (2.13) shown in Figure 2.6. The empirical universal coefficient $\alpha$ for (2.14) is also suggested as $\alpha_{U}=0.06$ for any type of antennas. It is important to note that the maximum deviation of gain reduction factor from the coefficient $\alpha_{U}$ can be converted into 0.5 dB in power level. The complete form of the proposed Friis formula can use the modified frontside or adjusted far-field gain such that:

$$
\begin{align*}
& S_{21}(R)=\frac{\lambda^{2} G_{t} G_{r}}{16 \pi^{2} R^{2}}\left(1-\alpha_{U}\left(\frac{R}{2 \lambda G_{F S t} / \pi^{2}}\right)^{-2}\right)\left(1-\alpha_{U}\left(\frac{R}{2 \lambda G_{F S r} / \pi^{2}}\right)^{-2}\right)  \tag{2.15}\\
& \times\left(1-\left|\Gamma_{t}\right|^{2}\right)\left(1-\left|\Gamma_{r}\right|^{2}\right)\left|\hat{\rho}_{t} \cdot \hat{\rho}_{r}\right|^{2} \\
& S_{21}(R)=\frac{\lambda^{2} G_{t} G_{r}}{16 \pi^{2} R^{2}}\left(1-\alpha_{U}\left(\frac{R}{2 \lambda G_{A t} / \pi^{2}}\right)^{-2}\right)\left(1-\alpha_{U}\left(\frac{R}{2 \lambda G_{A r} / \pi^{2}}\right)^{-2}\right)  \tag{2.16}\\
& \times\left(1-\left|\Gamma_{t}\right|^{2}\right)\left(1-\left|\Gamma_{r}\right|^{2}\right)\left|\hat{\rho}_{t} \cdot \hat{\rho}_{r}\right|^{2}
\end{align*}
$$

where all $\alpha$ are chosen to be $\alpha_{U}=0.06$. It is assumed that there is no losses from polarization or impedance mismatch so that $\Gamma=0$ and $\left|\hat{\rho}_{t} \cdot \hat{\rho}_{r}\right|=1$. The validity of the
universal coefficient will be discussed again when evaluating the power transmission in chapter 4.

## 2. 2. 3 Radiating Near-field Pattern Approximation

In previous sections, the term involving the universal gain reduction factor is used to calculate the boresight power transmission. In this section, another modifying term to the far-field pattern is presented to approximate the angular pattern variation within the Fresnel region.

The matching $\cos ^{q}(\theta)$ patterns to the far-field and Fresnel field pattern are employed in its derivation. Assuming the matching $q$-factors are $q$ and $q_{F}$, respectively, they can be calculated using


Figure 2.8: Approximation of the field pattern within Fresnel region using cosine pattern.


Figure 2.9: Simulation setup for the Fresnel field using waveguide such that the horn antenna is typically located at the far-field region of the waveguide.

$$
\begin{gather*}
G=2(2 q+1)  \tag{2.17}\\
G_{F}(\Delta)=G \cdot \gamma(\Delta)=2\left(2 q_{F}+1\right) \tag{2.18}
\end{gather*}
$$

Each matching $\cos ^{q}(\theta)$ pattern is multiplied by the corresponding gain such as the farfield gain or the gain in the Fresnel region. Then, the difference between two approximate matching $\cos ^{q}(\theta)$ patterns is defined as the modifying term,

$$
\begin{equation*}
M(\theta)=G_{F} \cdot \cos ^{2 q_{F}}(\theta)-G \cdot \cos ^{2 q}(\theta) \tag{2.19}
\end{equation*}
$$

By applying the modifying term to the original far-field pattern, the cosine approximation of the Fresnel field pattern with angular dependence can be obtained. A complete form of the Fresnel field pattern can be presented as

$$
\begin{equation*}
G_{F}(\theta)=G(\theta)+\left(G \cdot \gamma \cdot \cos ^{2 q_{F}}(\theta)-G \cdot \cos ^{2 q}(\theta)\right) \tag{2.20}
\end{equation*}
$$

where $\gamma=1-\alpha_{U}(R / 2 \lambda G)^{-2}$.

Here, $G$ inside the gain reduction factor $\gamma$ can be replaced by either $G_{F S}$ or $G_{A}$ to construct the universal gain reduction factor. To approximate the Fresnel field pattern, the information required is: 1) $2 D$ far-field pattern of the antenna, 2) operating frequency, and 3) separation distance $R$ between $T x$ and $R x$ antennas.

The near-field pattern of a rectangular waveguide operating at 12.7 GHz is simulated
using FEKO and compared with the result obtained using (2.20). Two separation distances within the Fresnel region $R=4 \lambda$ and $8 \lambda$ are investigated. Figure 2.8 shows the calculated and simulated radiation patterns in the Fresnel region, and a good agreement between those patterns is observed. As shown in Figure 2.9, both horn and waveguide are vertically polarized, and one waveguide is fixed and the other horn rotates at different angles. Using this simple approximation for the Fresnel field pattern, various scenarios such as off-axis and rotational scenarios can be obtained.


Figure 2.10: Comparison of the power transmission at $\mathrm{R}=4 \lambda$ in rotational scenario among 1) FEKO simulation, 2) Cosine approximation, 3) Power transmission using farfield pattern.

Figure 2.10 shows evaluation for rotational scenarios with Ku-band standard horn antennas. The power transmission using the cosine approximation is in a good agreement with the simulated power transmission using full-wave simulations.

### 2.3 Integral Form of the Coupling Formula

The integral form of coupling is aimed at computing mutual coupling between two antennas by employing the integral of the scalar product between two antenna's normalized vector far-field patterns. It is demonstrated in [13] that the mutual coupling can be efficiently obtained through the transmission integral [14], involving simple farfield patterns of two antennas and a separation distance between them. In recent years, there has been an advancement to allow the computation of the mutual coupling in the flexible movement [16], and its application extends from the large reflectors to the electrically small antennas for the RFID system [17]. In those works, the integral coupling offers the accurate coupling estimation within a relatively flexible antenna configuration. However, 3D radiation pattern and a relatively complex coupling program are often required, and a considerable computing may be requested for a large antenna.

With the coupling integral formula, it is possible to compute the coupling coefficient between two arbitrary antennas that are placed within the other's near-field region. This integral form of the coupling formula has been widely investigated, and the coupling integral provides good approximation for most of the antennas by using normalized vector far-field pattern of the antenna [13-17]. The amplitude of input wave to transmitting and output wave from receiving antenna are defined as $a_{T}$ and $b_{R}$, respectively. The coupling coefficient between the transmitting and receiving antennas is presented as [13]. A close integral form of the inner product between two vector electric fields can be used for computing the coupling coefficient between the transmitting and receiving antennas [13]. Note that the time convention, $e^{j w t}$ is utilized in the coupling
formula. The coupling coefficient between two transmitting and receiving antennas can be written as [13]

$$
\begin{equation*}
\frac{b_{R}}{a_{T}}=-\frac{C}{k} \iint_{\sqrt{k_{x}{ }^{2}+k_{y}{ }^{2}}<k} \frac{\bar{f}_{R}(-\bar{k}) \cdot \bar{f}_{T}(\bar{k})}{k_{z}} e^{-j \bar{k} \cdot \bar{r}} d k_{x} d k_{y} \tag{2.21}
\end{equation*}
$$

where $\bar{f}_{R}$ and $\bar{f}_{T}$ are the normalized vector far-field patterns for the receiving and transmitting antennas, respectively. $\bar{k}=\hat{x} k_{x}+\hat{y} k_{y}+\hat{z} k_{z}$ is the propagation vector. $\bar{r}$ is the vector from origin (usually phase center of the antenna) of the transmitting antenna to one of the receiving antenna. The integral over the transverse differential $d k_{x} d k_{y}$ makes it possible to remain only radiating part of the spectrum. The condition $\sqrt{k_{x}{ }^{2}+k_{y}{ }^{2}}<k$ allows the computation outside the reactive near-field region. In the near-field measurement [15], the similar boundary condition has been used. $C$ is defined as a mismatch constant,

$$
\begin{equation*}
C=\frac{Z_{\text {Feed }}}{\eta_{0}\left(1-\Gamma_{R} \Gamma_{L}\right)} \tag{2.22}
\end{equation*}
$$

Where $\eta_{0}$ is the intrinsic impedance, $Z_{\text {Feed }}$ is the characteristic impedance of the waveguide to feed the receiving antenna, and $\Gamma_{R}$ and $\Gamma_{L}$ are the antenna and load impedance reflection coefficients, respectively. The coupling in the radiating near field can be obtained with the information of the far-field patterns. It means that the integral coefficient is effective to obtain the coupling coefficients between antennas placed in the radiating near field.

## 2. 4 Evaluation of the Formula and Measurement

In this chapter, two formulas discussed above: 1) Friis formula with correction term and
2) Integral coupling formula are validated with various type of antennas placed within Fresnel region or closer radiating near-field region. For the Friis formula, the formula is evaluated by comparing it with the Chu's formula [7]. In addition, the applicability of the formula to the circularly polarized antenna is also validated. The full-wave simulation, FEKO is used in the evaluation. FEKO is advantageous since the size of the simulation does not depend on the separation distance but antenna configuration itself. Furthermore, using an advanced algorithm, the FEKO can efficiently reduce the number of interaction between triangular basis functions, which is named as MFMM (Multilevel Fast Multipole Method). For instance, normal MOM (Method of Moment) treats $N$ basis functions with $N^{2}$ memory usage and $N^{3}$ in CPU time. For the case of MFMM, it accommodates $N \times \log (N)$ scaling for memory and $N \times \log (N) \times \log (N)$ in CPU time. These features make it possible simulate electrically huge antenna configurations in various antenna placements. To validate this numerical evaluation, indoor measurement using standard gain horn is performed. All evaluations performed confirm the effectiveness of the proposed formula.

## 2. 4. 1 Numerical Evaluation of the Friis formula

The Friis formula with correction term is validated using full-wave simulation, FEKO. To verify the validity of the Friis formula presented in (2.15) and (2.16), a variety of antennas with a wide range of gains and operating frequencies are selected as listed in Table 2.1. The separation distance between the Tx and Rx antennas $R$ is varied in the antenna boresight direction to cover both Fresnel and far-field regions, and the power transmission levels are simulated using full-wave simulation tool FEKO and compared to the calculated results using the proposed formulas (2.15) and (2.16). Note that
combinations of both high-gain $(G>10 \mathrm{~dB})$ and low-gain $(G<10 \mathrm{~dB})$ antennas are carefully chosen to fully evaluate the applicability of (2.16).

Table 2.1: Variety of antennas used in the evaluation of the Friis formula.

| Group of <br> antennas | Antenna type | Far-field gain <br> $(\mathrm{G})$ | Front-side <br> Gain $\left(\mathrm{G}_{\mathrm{F}}\right)$ | Adjusted-side <br> Gain $\left(\mathrm{G}_{\mathrm{A}}\right)$ |
| :---: | :---: | :---: | :---: | :---: |
|  | Ku-band horn | 15.1 dB | G | G |
|  | C-band reflector <br> High-gain to | C-gin <br> (Circular horn feed) <br> C-band reflector <br> (Choked horn feed) | 49.5 dB | G |
| Low to Low | 1.4 GHz dipole | 2.14 dB | 2 G | G |
| High-gain to <br> Low-gain | C-band horn (TX) <br> C-band waveguide <br> $(\mathrm{RX})$ | 10.5 dB | 10.7 dB | G |

## A. High-Gain to High-Gain Antennas

The Ku-band standard gain horn and C-band reflector antenna are chosen as representative high-gain antennas to verify the formula. The boresight power transmission between the two antennas are simulated and compared to the results from


Figure 2.11: Comparison of the boresight power transmission (Ku-band horns) using the standard Friis formula, the asymptotic Friis formula, and simulation.


Figure 2.12: Comparison of the boresight power transmission (C-band reflector antennas) using the standard Friis formula, the asymptotic Friis formula, and simulation.
the proposed formulas (2.15) and (2.16). As suggested in Table 2.1, the adjusted gain for high-gain antennas are considered as same as the far-field gain of the antenna. The simulation results are shown in Figure 2.11 and Figure 2.12. An excellent agreement can be clearly observed.


Figure 2.13: Comparison of the boresight power transmission (1.4 GHz dipole antennas) using the standard Friis formula, the asymptotic Friis formula, and simulation.

## B. Low-Gain to Low-Gain Antennas

The dipole antennas operating at 1.4 GHz is used to validate the low-gain antennas case. For a dipole antenna, the modified gain, $G_{F S}$ can be calculated as $2 G$, due to the use of one-side total radiated power. Similarly, $2 G$ can be substituted as the adjusted gain, $G_{A}$, which is applicable to low-gain antennas. Comparison of the both formulas to the simulated power transmission is presented in Figure 2.13. In the Fresnel region the
proposed formula shows a satisfactory match to the simulated power transmission, and an enhancement of more than 5 dB , which is equivalent to $60 \%$ deviation of the power transmission, by the proposed formula compared to the Friis formula is well appreciated.


Figure 2.14: Comparison of the boresight power transmission (waveguide and horn antennas) using the standard Friis formula, the asymptotic Friis formula, and simulation.

## C. High-Gain to Low-Gain Antennas

The C-band rectangular horn and rectangular open-ended waveguide are chosen as a transmitting high-gain and receiving low-gain antennas. Again a good agreement is observed in Figure 2.14. It is concluded that the proposed formula is also applicable to two different antennas from different groups.

Therefore, the information required to calculate the power transmission using the
proposed formula is based on three parameters that may be easily accessible: 1) far-field gain at boresight, 2) operating frequency, and 3) separation distance. Non complex 3D

Table 2.2: Nearest range of the valid prediction.

| Antenna type | Ku-band horn | C-band reflector | 1.4 GHz dipole | C-band horn and <br> waveguide |
| :---: | :---: | :---: | :---: | :---: |
| Nearest effective <br> distance | $0.18 \times\left(2 \mathrm{D}^{2} / \lambda\right)$ | $0.25 \times\left(2 \mathrm{D}^{2} / \lambda\right)$ | $0.35 \lambda$ | $2.5 \lambda$ |

patterns or calculations are involved. The calculated results show good agreement with the simulated power transmission. It must be noted that the effective range of the proposed formula is limited to the Fresnel region. One can observe that all of the power transmissions from the formula decrease drastically after the nearest distance of the Fresnel region. The failure of the formula attributes to the quadratic form of the formula, which becomes negative at a very close distance. The nearest valid ranges for various antennas are summarized in Table 2.2, which indicates the distance in terms of far-field edge distance or wavelength. For high-gain antennas, the formula can predict the accurate power transmission as close as $0.2 \times\left(2 D^{2} / \lambda\right)$. The nearest distance for dipole is similar to the range that the integral coupling formula [17] predicts, which is approximately $0.35 \lambda$.

## 2. 4. 2 The Chu's Formula: Comparison to the Friis Formula

An approximate generalized Friis formula was proposed by Chu [7] which is based on the response of the fundamental Gaussian mode:

$$
\begin{equation*}
S_{21}=\frac{P_{r}}{P_{t}}=\frac{A_{t} A_{r}}{\left(\frac{A_{t}+A_{r}}{2}\right)^{2}+\lambda^{2} R^{2}} \tag{2.23}
\end{equation*}
$$

Using (2.10) it can be rewritten as

$$
\begin{equation*}
S_{21}=\frac{P_{r}}{P_{t}}=\frac{\lambda^{2} G_{t} G_{r}}{16 \pi^{2} R^{2}} \frac{1}{1+\frac{\lambda^{2}}{16 \pi^{2} R^{2}}\left(\frac{G_{t}+G_{r}}{2}\right)^{2}} \tag{2.24}
\end{equation*}
$$

It is now interesting to see if (2.24) can be cast in the format of (2.4) for the case when $G_{t}$ $=G_{r}=G$. If one applies following,

$$
\begin{equation*}
\frac{1}{1+x} \approx 1-x \sim\left(1-\frac{1}{2} x\right)^{2} \tag{2.25}
\end{equation*}
$$



Figure 2.15: Chu's formula in the near-field region without any failure in very close distances.

Approximation assuming that $x$ is smaller than one, then (2.24) can be recast to the following equation,

$$
\begin{equation*}
S_{21}=\frac{P_{r}}{P_{t}} \approx \frac{\lambda^{2} G^{2}}{16 \pi^{2} R^{2}}\left[1-\frac{\pi^{2}}{128}\left(\frac{\pi^{2} R}{2 \lambda G}\right)^{-2}\right]^{2} \tag{2.26}
\end{equation*}
$$

Comparing (2.26) with (2.4) and (2.6), it is concluded that Chu's Friis formula can also
be approximately expressed by the classic Friis formula and a quadratic correction term with the same normalized distance defined in (2.6). A universal coefficient $\alpha$ of $\pi^{2} / 128$ is implied, which is close to the empirical universal coefficient $\alpha$ of 0.06 for aperture antennas obtained through case studies. The main advantage of the Chu's formula is its effectiveness within the very close radiating near-field region. Figure 2.15 shows that the formula can estimate without any failure in the close near-field region. However, it is


Figure 2.16: Chu's formula in the near-field region and its failure for small gain antennas.
worth mentioning that the formula (2.44) is still not effective for the small gain antenna case such as dipole antenna. Figure 2.16 implies that the formula rarely provide an enhancement to the Friis formula for small gain antennas such as dipole antennas.

## 2. 4. 3 Numerical Evaluation of the Integral Coupling Formula

The integral coupling formula is useful to compute the cross-talk between two antennas closely spaced. For instance, multiple large reflector antennas can be placed very closely
for satellite communication and deep-space applications. The inter-spacing for a electrically large antennas is equivalent to a very close distance in the radiating nearfield, respective to the dimension of the antenna. Therefore, an accurate estimation for a very close distance is required. In this aspect, the integral coupling formula is advantageous, which makes it possible to estimate the coupling within 50 cm for $2-3 \mathrm{~m}$ diameter reflector antennas.


Figure 2.17: Coupling between two C-band reflectors placed in a close distance.


Figure 2.18: Coupling between two C-band reflector antennas.

Figure 2.17 depicts the configuration of the 1 m diameter reflector antenna. Two primefocus reflector antennas with diameters of 1.9 m and 2 m are simulated with FEKO. The operating frequency is 5.5 GHz . The simulated directivities are 39.62 dB and 39.45 dB , respectively. The simulated patterns are then used in the computer program and the coupling between the reflector antennas is calculated versus separation distances. The calculated coupling values are compared with FEKO simulations and the Friis formula, as shown in Figure 2.18. An excellent agreement between results of the integral coupling formula and FEKO results can be observed. In the near-field region, the coupling program shows significant improvement, compared to the Friis formula.


Figure 2.19: Coupling between two closely spaced reflector antennas.

Another evaluation is the closely spaced reflector antennas. As discussed above, in the modern communication system, the cross-talk between two independent communication systems is a critical problem. Using the coupling program, the cross-talk between two antennas can be accurately estimated. Especially, it is not easy to estimate the cross-talk
between large structures such as reflector antennas. The coupling program maybe the optimum solution to analyze this problem. The geometry of the two offset reflectors is shown in Figure. 2.19. Two reflectors are tilted and faces toward different directions. Using the simulated far-field pattern, the coupling level is estimated, and compared with the FEKO simulation results. The coupling program provides -80.81 dB S21 and the coupling program calculates the coupling as -77.14 . If we consider very small signal level, the results are very similar. It is proven that the coupling program is also effective in the arbitrary rotating geometries.

## 2. 4. 4 Linkbudget Analysis between Circularly Polarized Antennas

In this chapter, near-field linkbudget between two antennas which are diversely polarized is evaluated using full-wave simulation and the generalized Friis formulas presented in Chapter 2.3. The circularly polarized antennas are widely used due to its advantages such as robustness against the fading and polarization misalignment. In spite of the popular usage, there has been no studies on the near-field linkbudget between two circularly polarized antennas. The main difference in the formulas which uses circularly polarized antenna is the polarization mismatch. Compared to the linearly polarized antenna, the expression of the polarization mismatch becomes slightly complicated. The loss caused by the circular polarization mismatch is given in [18], and this expression can be incorporated into the generalized Friis formula (2.15).

The polarization mismatch defined in [18] is used in this dissertation. The polarization efficiency can be written as

Polarization efficiency $=\frac{\left|\bar{E}_{1} \cdot \bar{E}_{2}^{*}\right|^{2}}{\left|\bar{E}_{1}\right|^{2}\left|\bar{E}_{2}\right|^{2}}=\frac{1+\left|\sigma_{1}\right|^{2}\left|\sigma_{2}\right|^{2}+2\left|\sigma_{1}\right|\left|\sigma_{2}\right| \cos \left(\delta_{1}-\delta_{2}\right)}{\left(1+\left|\sigma_{1}\right|^{2}\right)\left(1+\left|\sigma_{2}\right|^{2}\right)}$
where $\delta 1$ and $\delta 2$ are the phases of the polarization ratios of the magnitudes of the right handed and left handed components $E_{R}$ and $E_{L}$ for the antenna 1 and antenna 2. The electric field can be written using right-handed and left-handed orthogonal components $E_{R}$ and $E_{L}$. The parameter $\sigma$ is the ratio between two components. The coupling between diversely polarized antennas is evaluated using full-wave simulation and the generalized Friis formula which includes the polarization mismatch loss addressed above. The complete generalized Friis formula used in this evaluation can be written as

$$
\begin{align*}
S_{21}(R) & =\frac{\lambda^{2} G_{t} G_{r}}{16 \pi^{2} R^{2}}\left(1-\alpha_{U}\left(\frac{R}{2 \lambda G / \pi^{2}}\right)^{-2}\right)\left(1-\alpha_{U}\left(\frac{R}{2 \lambda G / \pi^{2}}\right)^{-2}\right) \\
& \times\left(1-\left|\Gamma_{t}\right|^{2}\right)\left(1-\left|\Gamma_{r}\right|^{2}\right) \times\left.\hat{\rho}_{t} \cdot \hat{\rho}_{r}\right|^{2}= \\
= & \frac{\lambda^{2} G_{t} G_{r}}{16 \pi^{2} R^{2}}\left(1-\alpha_{U}\left(\frac{R}{2 \lambda G / \pi^{2}}\right)^{-2}\right)\left(1-\alpha_{U}\left(\frac{R}{2 \lambda G / \pi^{2}}\right)^{-2}\right)  \tag{2.28}\\
& \times\left(1-\left|\Gamma_{t}\right|^{2}\right)\left(1-\left|\Gamma_{r}\right|^{2}\right) \times \frac{1+\left|\hat{\sigma}_{1}\right|^{2}\left|\hat{\sigma}_{2}\right|^{2}+2\left|\hat{\sigma}_{1}\right|\left|\hat{\sigma}_{2}\right| \cos \left(\delta_{1}-\delta_{2}\right)}{\left(1+\left|\hat{\sigma}_{1}\right|^{2}\right)\left(1+\left|\hat{\sigma}_{2}\right|^{2}\right)}
\end{align*}
$$

The circularly polarized horn antenna that will be discussed in Chapter 5 is employed in this evaluation. The configuration of the horn antenna is shown in Figure 2.20.


Figure 2.20: The horn antennas used in the evaluation: (a) horn antenna with smooth septum polarizer, (b) Linearly polarized standard horn antenna.

There are two horn antennas: 1) Horn antenna with a smoothly contoured circular polarizer (antenna I), 2) Linearly polarized horn antenna (antenna III). The evaluation is performed at the antenna operating frequency of 5.8 GHz . Each antenna parameters follow as below:

- Circularly polarized (CP) horn antenna: Directivity (CP) at $5.8 \mathrm{GHz}=11.2 \mathrm{~dB}$, Impedance matching at $5.8 \mathrm{GHz}:-20 \mathrm{~dB}$, Axial ratio $(\mathrm{AR})=0.44 \mathrm{~dB}, \mathrm{D}$ (diameter
of the aperture $)=1.3 \lambda$.
- Linearly polarized (LP) horn antenna: Directivity at $5.8 \mathrm{GHz}=11.28 \mathrm{~dB}$, Impedance matching at $5.8 \mathrm{GHz}:-15.8 \mathrm{~dB}$, Axial ratio $(\mathrm{AR})=\infty$ (actually 55 dB ), D (diameter of the aperture $)=1.3 \lambda$.

The two meaningful scenarios are evaluated: 1) circularly polarized horn to circularly polarized horn antennas, 2) linearly polarized horn to circularly polarized horn antenna. The evaluated scenarios are shown in Figure 2.21. The far-field distance of both horn antennas is calculated as roughly $3.5 \lambda$. Based on this, the separation distance $(R)$ is varied from $1.2 \lambda$ to $4.0 \lambda$ so as to evaluate the CP linkbudget in the Fresnel region. It was assumed that the phase center of the antenna is placed on the aperture of the antenna. The scenarios used in the evaluation is depicted in Figure 2.21. The power transmission level using the full-wave simulation and the generalized Friis formula is shown in Figure 2.22 and Figure 2.23. For the case of the CP horn to CP horn, there is no polarization mismatch loss since the same CP antennas were employed. A good agreement can be observed between the simulation and the generalized Friis formula for this case. For the case of CP antenna to LP antenna, it is interesting to observe the change in the linkbudget since all parameters of the CP and LP antennas are very similar except for the polarization difference. The polarization mismatch loss between CP and LP can be calculated using the equation (2.27). Since we can use the relationship, $\sigma=\frac{A R-1}{A R+1}$, we can derive $\sigma$ (linear polarization $)=1$ and $\sigma$ (circular polarization $)=0$. If we substitute this parameters into the equation (2.27), the polarization power loss becomes 0.5 as expected.

The coupling level shown in Figure 2.23 is around 3dB below the coupling level presented in Figure 2.22. For this case, there is a good agreement between simulated results and generalized Friis formula.


Figure 2.21: the near-field coupling simulation between circularly polarized horn and the linearly polarized horn antennas.


Figure 2.22: Linkbudget between two CP horn antennas.


Figure 2.23: Linkbudget between CP and LP horn antennas.

### 2.4.5 Measurement Results

With the development of computational capabilities, it becomes possible to obtain the power transmission levels between large antennas with complex configurations using
full-wave simulations. A comprehensive evaluation is performed through both full-wave simulations and measurements utilizing two Ku-band standard gain horn antennas.

Two identical Ku-band standard gain horns are employed as the transmitting and receiving antennas to evaluate the Friis formula. A single operating frequency of 12.7 GHz is chosen. The aperture size of the horn is $50.8 \mathrm{~mm} \times 37.6 \mathrm{~mm}(2.15 \lambda \times 1.59 \lambda$ at 12.7 GHz). The relatively small wavelength makes in-door far-field measurements feasible and accurate measurement results can be obtained with the available measurement equipments.

(a)

(b)

Figure 2.24: Setup of the (a) full-wave simulation using FEKO and (b) in-door measurement.

The separation distance $R$ is measured between the antenna phase centers. The setup of the simulation and measurement is illustrated in Figure 2.24. The Tx and Rx horns are mounted at the same height on an optical table, and aligned toward each other in the boresight direction. The position of the Tx horn is fixed, and the Rx horn is moved along its boresight. The separation distance between the antenna phase centers is $R$. Microwave absorbers are used to cover the exposed metal surfaces in order to effectively reduce reflection and diffraction effects caused by the optical table and other supporting structures. Both horns are vertically polarized. Any undesired losses caused by polarization mismatch, therefore, can be avoided. The return loss of both horns is small enough so that the input impedance mismatch is negligible. A vector network analyzer is employed to measure the $S_{21}$, the power transmission level between them.

Computer simulations are also implemented using FEKO. It is a powerful full-wave simulation tool based on the method of moments (MoM). The complexity of FEKO analysis is primarily determined by the antennas' electrical sizes, but almost not affected
by the separation distance between them. This unique property makes FEKO more suitable for efficiently analyzing the interaction between antennas in the far-field region. An accurate and efficient analysis of this type of relatively large structures can be achieved.

First the far-field radiation pattern of the Ku-band standard gain horn is simulated using FEKO. It is worth pointing out that the standard gain horn has a specially designed curved flare to help smooth the phase distribution at the horn aperture and improve the gain performance at the desired frequency band, but in FEKO simulation a simplified model with straight flare is used. To verify the validity of the computer model used in FEKO simulations, the Ku-band standard gain horn is measured in the UCLA spherical near-field range.


Figure 2.25: The Ku-band standard gain horn is measured in the UCLA spherical near-field range.


Figure 2.26: Normalized far-field radiation patterns (H-plane) obtained by FEKO simulation and measurement.

Figure 2.25 shows the near-field measurement setup. The Ku-band standard gain horn is mounted on a positioner, and by rotating the horn and the positioner, the amplitude and phase of the radiated near field over an enclosed spherical surface are obtained. The farfield radiation pattern can then be accurately achieved using the measured near-field data. The measured far-field radiation pattern is then used as a reference and compared with the simulated far-field pattern using FEKO, as shown in Figure 2.26. The simulated and measured gains are 15.28 dB and 15.47 dB , respectively. An excellent overall agreement can be observed. In the backlobe region measurement predicts lower levels compared with simulation results because that part of the radiation energy is constantly blocked by the positioner throughout the entire measurement procedure (see Figure 2.24). The simulated gain is 0.19 dB lower than the measurement, which can be attributed to the simplified straight flare used in the FEKO model. Note that for this type of antenna the
directivity and gain are very close considering that the conducting and other losses are negligible.

The separation distance $R$ is varied from $4 \lambda$ to $31 \lambda$ in the antenna boresight direction, and the power transmission levels between the Tx and Rx horns are simulated and measured. The estimated $S_{21}$ using the Friis formula is also calculated where the measured gain of 15.47 dB is used. Note that the edge of the far-field region is estimated to be $14.3 \lambda\left(2 D^{2} / \lambda\right.$, where $D$ is the diameter of the sphere to enclose the antenna). This test range, therefore, well covers both Fresnel and far-field regions.


Figure 2.27: Boresight transmission levels between the Tx and Rx horns. The edge of the far-field region $\left(2 D^{2} / \lambda\right)$ is $14.3 \lambda$.

Figure 2.27 compares the boresight transmission levels between the horns. It can be observed that the FEKO simulations agree extremely well with the indoor measurements. In the far-field $(R=29 \lambda)$, the simulated $S_{21}$ is 0.4 dB lower than the measurement, which can be predicted using (2.1) if considering the 0.19 dB difference between the simulated and measured gains of the Ku-band horn. The Friis formula provides very accurate estimations in the far-field region. Compared with the measurement the difference between these two curves approaches its minimum of less than 0.1 dB at $R=31 \lambda$. However, the performance of the Friis formula deteriorates as the separation distance moves into the Fresnel region. At the edge of the far-field region $(R=14.3 \lambda)$ the estimated $S_{21}$ using the Friis formula is 0.3 dB more than the measurement. A maximum deviation of 1.9 dB is observed at the closet separation distance $(R=4 \lambda)$ investigated in measurements.

Therefore, a good knowledge of the antenna gains is crucial to accurate estimations using the Friis formula. It is expected because, according to (2.2), the power transmission level is proportional to $G_{t} G_{r}$ at a fixed separation distance in terms of wavelength. The Friis formula is able to accurately estimate the power transmission levels in the far-field region, but its accuracy deteriorates considerably in the Fresnel region.

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## Chapter 3

## RF MEMS PASS Antenna

The frequency reconfigurable patch antenna with switchable slot (PASS) is achieved with a commercially packaged MEMS (microelectromechanical system) switch, and a bias network is designed to separate the RF and DC signals. It is difficult to operate the MEMS switch placed in the slot structure [1-2]. In this dissertation, the fully integrated dc bias network is presented, which is capable of actuating the MEMS switch on the slot. The bias network functions without adverse effect on dc bias and RF radiation mechanism. A negligible loss is generated by including the bias network. The switchable slot on the patch antenna can change the current path, which enables the patch antenna to operate at 4.57 GHz and 4.88 GHz . In this design, a commercially available MEMS switch is used. The switch is advantageous due to its low insertion loss and high isolation level [3-7]. The MEMS patch antenna shows reasonable return losses and directivities at the broadside direction for both switch on and off states. In addition, the effect of RF MEMS switch on the antenna impedance and radiation pattern will be discussed.

## 3. 1 Patch Antenna with Switchable Slot (PASS)

The PASS (Patch Antenna with Switchable Slot) has been widely used in space
exploration and commercial communication services [8]. The patch antenna can reconfigure the frequency bands or senses of the different circular polarizations. In this dissertation, we mainly focused on the frequency reconfigurable patch antenna. Figure 3.1 depicts the PASS design which can change the resonant frequency of the patch antenna. By switching on or off, shorter or longer current path is generated, which makes the patch antenna resonate at higher or lower frequency, respectively. By adjusting the length of the slot, the tuning range of the frequency reconfigurability can also be controlled.


Figure 3.1: The concept of the frequency reconfigurable PASS (Patch Antenna with Switchable Slot).

## 3. 2 RF MEMS PASS Antenna Design

In the recent past, the PASS design was implemented with the diode switches [8]. The diode switch is easy to implement the reconfigurable antenna owing to its two pin structure, however, it shows a high insertion loss at RF frequency bands. As mentioned
above, the MEMS switch is better in terms of switch loss, but three pin commercial MEMS switch needs an additional DC line, which makes the bias network complicated. In this dissertation, an innovative dc bias line is integrated into slot-loaded structure to actuate the MEMS in the slot. By including dc bias line in the design, the design accommodates easy supply of dc potential to the RF switch and full integration on one layer. Additionally, the bias network consists of the spiral and interdigital gap filters that are utilized to minimize the interference. The proposed bias network is applicable to any other slot loaded structure for commercial three terminal MEMS switches.

## 3. 2. 1 PASS Antenna Design

The slot-loaded patch antennas exhibit two resonant frequencies shifting from 4.57 GHz to 4.88 GHz , according to the direct or meandering current paths of the slot. The topology of the proposed PASS antenna co-designed with a packaged RF MEMS switch is shown in Figure 3.2. The patch antenna with a slot is printed on the substrate, and SMA probe is connected to the patch antenna for the excitation. The position of the probe feed is placed at the position, $(\mathrm{x}, \mathrm{y})=(4.8 \mathrm{~mm}, 0)$ for the best match to $50 \Omega$. The size of the patch antenna ( $\mathrm{L} \times \mathrm{W}$ ) is $17.7 \mathrm{~mm} \times 20.5 \mathrm{~mm}$ and the slot has the dimension (Ls x Ws) of $19.6 \mathrm{~mm} \times 2.2 \mathrm{~mm}$ with $\mathrm{Wg}=1 \mathrm{~mm}$. The substrate used in this antenna is Rodgers Duroid


Figure 3.2: Patch antenna with switchable slot (PASS), bias network and RF MEMS switch: Top and side view of the PASS antenna configuration.

5880, which has thickness of 3.18 mm , size of $55 \mathrm{~mm} \times 55 \mathrm{~mm}$, and relative dielectric constant $\varepsilon \mathrm{r}=2.2$ with a loss $\tan \delta=0.0009$ at 5 GHz . The RF MEMS switch placed in the slot alters effective electrical lengths of the patch antenna. When the switch is at the offstate, the current meanders around the slot which result in the resonance at 4.57 GHz . The switch at the on-state creates the direct current path across the slot for operating at 4.88GHz. The RF MEMS switch and bias circuit for the actuation of the switch will be discussed next.


Figure 3.3: Radant MEMS SPST-RMSW 100 packaged switch and description of MEMS switch: source, drain, and gate.

## 3. 2. 2 RF MEMS Switch and Integrated Circuit Components

Radant MEMS SPST-RMSW 100 packaged switch is used in the PASS antenna as shown in Figure 3.3. The switch is placed in the center of the slot. The copper pad ( 1.5 mm X 1.5 mm ) is used to attach the gold base of the MEMS switch. The actuation voltage 90 VDC between the gate and source is applied to the pad in the bias line. When the switch is at the on-state, the RF continuity is generated with approximately 0.23 dB insertion loss. In the off-state, 20 dB isolation between source and drain creates RF discontinuity. To guarantee proper operation, source and drain pads are connected to the microstrip antenna with two wire-bondings. This method can reduce the relatively high-impedance of the wire-bonding line and increase the current flow. The width of the bias line is 0.2 mm which results in high characteristic impedance $\mathrm{Z}_{\mathrm{line}}>200 \Omega$ at the operating frequencies of the antenna from 4.6 GHz to 5 GHz . To actuate the switch, the source and drain of the MEMS switch is required to connect to the dc ground plane. The RF ground plane is utilized for the dc ground plane for the simplicity of the circuitry. The dc


Figure 3.4: The dimension and specification of the bias network: bias network consists of the dc bias line, spiral filter and interdigital gap filter.

Table 3.1: Dimension of the PASS patch antenna

| $\mathrm{L}^{2}$ | 17.70 mm | $\mathrm{~L}_{5}$ | 0.55 mm | $\mathrm{~L}_{10}$ | 0.10 mm | $\mathrm{~W}_{1}$ | 0.30 mm |
| :---: | ---: | :--- | :--- | :--- | :--- | :--- | :--- |
| $\mathrm{~L}_{1}$ | 3.30 mm | $\mathrm{~L}_{6}$ | 1.90 mm | $\mathrm{~L}_{11}$ | 1.20 mm | $\mathrm{~W}_{2}$ | 0.26 mm |
| $\mathrm{~L}_{2}$ | 2.80 mm | $\mathrm{~L}_{7}$ | 0.50 mm | $\mathrm{~L}_{12}$ | 2.30 mm | Wo | 3.85 mm |
| $\mathrm{~L}_{3}$ | 0.65 mm | $\mathrm{~L}_{8}$ | 0.50 mm | $\mathrm{~L}_{\mathrm{b}}$ | 19.00 mm | $\mathrm{G}_{1}$ | 0.08 mm |
| $\mathrm{~L}_{4}$ | 0.55 mm | $\mathrm{~L}_{9}$ | 1.85 mm | W | 22.10 mm | $\mathrm{G}_{2}$ | 0.30 mm |

continuity between patch antenna and RF ground plane is realized by utilizing the $\lambda / 4$
length stub line short to the RF ground plane through a via structure. The stub is placed at 4.4 mm above from the center of the patch antenna. The shorted line appears as open-state
for the antenna at the junction of the line and antenna due to its electrical length. The overall structure including the bias network is incorporated in a layer and only single via is used for dc ground. These features are amenable for the fully-integrated structure with minimum cost of fabrication.

(a) DC potential distribution

(b) Radiating current distribution

Figure 3.5: The configuration of the dc potential distributions and radiating current distributions of the bias network.

## 3. 2. 3 Operation of the Bias Network

The slot-loaded structure complicates implementing a proper bias network integrated in the antenna. This is because the microstrip patch around the slot limits the accessibility of the dc bias line. To increase the accessibility, the area that RF and dc signals coexist is placed around the slot as shown in Figure 3.4, and the specification is presented in Table 3.1. The design makes it feasible to maintain the high dc potential from the bias line to the gate of the switch and to continue the radiating current flow around the slot. The bias network is capable of strictly discerning the radiating and dc bias mechanism with the help of the high impedance filters in dc and microwave regions. The bias network includes the dc bias line, interdigital capacitor and spiral filter as shown in Figure 3.4. The interdigital capacitor [9] is widely used for many applications due to its high Qresponse as the low band stop filter. It also exhibits the low insertion loss in the pass band of the RF operation. In this design, the capacitor is used to isolate high dc potential from the microstrip patch and to flow the radiating current with small loss. Another important component in the bias network is the spiral filer [10]. The filter exhibits useful features like steep band rejection characteristics and compact size compared with widely used quarter-wave open stub. High impedance characteristics at the operating frequencies of the antenna prevent the leakage of the radiation current into the bias line. In Figure 3.5, the operating mechanism of the dc and radiating current is presented.


Figure 3.6: The operation of two filter components (spiral inductive filter and interdigital gap capacitive filter) of the bias network.

The dc potential is supplied along bias line without intruding other RF radiating regions. In the case of the radiating current, the interference with the bias line is minimized and its continuity on the patch is maintained. The S21characteristics of the single filter component is shown in Figure 3.6. The insertion loss of the interdigital capacitor is from 0.3 dB to 0.6 dB at the operating frequencies of the antenna from 4.6 GHz to 5.1 GHz . It exhibits the band rejection characteristics at the low frequency bands. The filter offers the sufficient band rejection characteristics against radiation current. The S21 of the inductor is lower than -12 dB from 4.57 GHz to 5.1 GHz , the range of dual-band operations of the antenna. Spiral filter should be placed in a position close to the patch to prevent standingwaves.

## 3. 3 Effects of the MEMS Switch on RF Performance

Incorporating the MEMS switch into the antenna design can shift the resonance of the antenna and change the radiation pattern [11]. Wire-bonding provides connection between switch and antenna. However, an additional inductance can be generated by applying wire-bonding, which can shift the resonance of the antenna. The MEMS switch itself involves the capacitance between the membrane and ground on the switch. Therefore, it is necessary to study robustness against the adverse effect. In order to investigate the effect of the MEMS switch, the PASS designs incorporated with MEMS model and ideal connection are simulated. The change of the performance with respect to the impedance matching and radiation pattern is investigated.

## 3. 3. 1 Simulated MEMS Switch and Ideal Connection

The original impedance and radiation pattern of the antenna without any switches can be altered when thin wire-bonding and switch is loaded with the antenna. To evaluate the changes caused by the wire-bondings and switch, the operation of the antenna loaded with RF MEMS switch should be compared with the antennas with an ideal connection. In Figure 3.7, two simulated cases when the switch is on or off are presented. For the case of the ideal connection, thin copper microstrip is placed across the middle of the slot to create ideal shortest path of the radiating current as shown in Figure 3.7 (a). The ideal


Figure 3.7: The simulated configurations for each case: a) ideal switch-off connection, b) wire-bonded MEMS switch-off, c) ideal switch-on connection, d) wire-bonded MEMS switch-on.
switch-off state is represented by placing the isolated copper pad as depicted in Figure 3.7 (c). In Figure 3.7 (b), the wire bonded RF MEMS switch model, which is closer to the real MEMS switch, is shown. The membrane of the MEMS switch, designed with simplified 0.075 mm width microstrip line, is placed on the silicon substrate with a height of 0.25 mm . Two gold wires with a 0.01 mm radius are used to connect the membrane and the patch antenna. On or off-state of the RF MEMS switch is represented by attaching or detaching the middle of membrane, respectively.

## 3. 3. 2 MEMS Switch and RF Performance

The S-parameters of the antenna for the each case are simulated and compared in HFSS.

All simulated configurations are identical except for using different kinds of connections across the slot: ideal connection and the wire-bonded RF MEMS switch. In Figure 3.8, the S-parameters with ideal and MEMS loaded connection are compared. At the off-state, the maximum resonance of the ideal and MEMS loaded connection occurs at 4.59 GHz and 4.62 GHz , respectively. At the on-state, the patch antenna operates at 4.89 GHz and 4.92 GHz with MEMS load and ideal connection, respectively. It is seen that a design loaded with a RF MEMS switch does not exhibit the significant shift of the resonant frequency, except for off-state. At the off-state, an additional capacitance of the MEMS switch might cause the shift of the center frequency by $30 \mathrm{MHz}(0.8 \%$ shift respect to the center frequency). Next, the change of the radiation pattern should be carefully considered to examine the loss of the radiation when the MEMS switch is loaded in the design. In Figure. 3.9, the simulated radiation patterns at frequencies of maximum resonances for each case are shown. The comparison was made with co-polarization and


Figure 3.8: the comparison of the return loss characteristics: 1) ideal connection, 2) wirebonded MEMS switch. a) switch-off state, b) switch-on state.
cross-polarization pattern at the off-state and on-state of the switch, respectively. The radiation patterns of MEMS switch are normalized to that of the ideal connection case. It is observed that the radiation patterns of three cases agree well with each other. The distortion caused by RF MEMS switch is still tolerable when the switch is in both off and on-state. At the broad-side direction, the loss of the co-polarized radiation pattern at switch off-state and on-state is less than 0.3 dB after loading the MEMS switch. Since all the cuts of the radiation pattern is not investigated, it is hard to generalize the change of the cross-polarization; however, from the Figure 3.9, roughly 1-3 dB rise in the crosspolarization is observed with a RF MEMS switch.


Figure 3.9: Simulated normalized co-polarized and cross-polarized radiation patterns at switch off-state and on-state for each case: 1) ideal connection, 2) wire-bonded MEMS switch.

## 3. 4 Fabrication and Measurement of the Prototype

The PASS antenna design is fabricated, and the impedance matching and radiation pattern of the design are measured [12]. The issues during the fabrication process are discussed. The measurement setup and measured results are provided in order to verify the numerical approach performed in the previous chapter. The measurement shows a good agreement with the simulated results.

## 3. 4. 1 Fabrication and Measurement Setup

The proposed antenna was fabricated and measured to demonstrate its operational characteristics. The photograph of the prototype antenna and the RF MEMS switch is shown in Figure 3.10. The presented design is carefully fabricated using etching facility in UCLA Center for High Frequency Electronics (CHFE). Main issue in the etching process is the accuracy of fabrication in the inter-digit gap design. The mask for the fabrication of the interdigital capacitor is designed for $80 \mu \mathrm{~m}$ gap, however, it is possible for the interdigital gap to be over-etched because $80 \mu \mathrm{~m}$ is around the limitation of the standard PCB fabrication. As shown in Fig. 3.11, over-etched case (worst case: 100um gap) for interdigital gap is simulated and it shows that insertion loss increases around 0.30.5 dB in antenna operation range as the slot gap increases. Therefore, it can be mentioned that the design is relatively robust against the loss during the etching process.

## 3. 4. 2 Measured Results

Using an Agilent 8720ES vector network analyzer, the return loss of the antenna was


Figure 3.10: The photograph of the fabricated design loaded with RF MEMS switch, dc

> bias line and bias circuitries.


Figure 3.11: The insertion loss of the inter-digit filter depending on the degree of the over-etching process: The frequency band of our interest is indicated using grey area.
measured. To observe the reconfigurability, the voltage supplier was also connected to the bias network while maintaining the connection of the RF cable connection to the network analyzer. The voltage 3VDC from the power supplier Agilent E3631A was amplified to 90VDC through an amplifier circuit. The actuation voltage 90VDC from the circuit was then applied to pad A through a soldered electrical wire as shown in Figure 3.12. Similarly, a pad B was connected to the ground of dc power supply to make sure that the patch was grounded. When the voltage was applied, the MEMS switch is actuated and created the short current path across the slot at 4.88 GHz . When no voltage was applied, the current around slot created the long current path at 4.57 GHz . The simulated and measured return loss of the antenna is shown in Figure 3.13. For more accuracy in simulation, the design was loaded with wire-bonded MEMS switch model as shown in Figure. 3.7 (b) and Figure 3.7 (d). The antenna resonated at 4.57 GHz at the switch-off state and 4.88 GHz at the switch-on state.


Figure 3.12: Measurement setup: 90 V voltage difference is applied to the both pads in the PASS antenna design. At the same time, the S-parameter of the antenna was measured.


Figure 3.13: Simulated and measured return loss of the proposed antenna.

There was a 30 MHz discrepancy between simulated and measured S-parameters. The over-etching of the interdigital capacitor is assumed to be the main reason for the mismatch. The interdigital gap was originally designed as $80 \mu \mathrm{~m}$, however, it is overetched to $100 \mu \mathrm{~m}$ gap in the prototype. Using HFSS simulation, it is found that additional $0.3-0.5 \mathrm{~dB}$ loss is generated from the over-etching. The measured frequency ratio of two operation states is $4.88 / 4.57=1.07$ and the separation between two states is 300 MHz . The radiation pattern was measured with the fabricated patch which includes ideal off-state and on-state connection. The measurement setup is shown in Figure 3.14. As discussed in section III, this model was verified to have almost similar return loss and radiation pattern to the case of actual MEMS switch. The normalized measured directivity pattern


Figure 3.14:The PASS antenna in the UCLA anechoic spherical near-field chamber.of the proposed antenna.
is shown in Figure 3.15. The directivity pattern was measured with ideal off-state and MEMS off-state design at 4.57 GHz and with ideal on-state design at 4.88 GHz . At the offstate, the measured patterns with ideal connection and MEMS switch are very similar. The measured directivities of the antenna at the off-state and on-state are 7.1 dBi and 7.6 dBi at the broadside direction, respectively. At the switch on-state, the crosspolarization is 20 dB lower than the co-polarization. However, at the switch of off-state, the cross-polarization becomes higher because the current meandering around the slot is relatively more sensitive to the asymmetrical structure of the bias network. The radiation pattern with cross-polarization can be enhanced with positioning bias network near the center of the antenna.


Figure 3.15: Measured normalized radiation pattern of the ideal off-state at 4.57 GHz and the ideal on-state at 4.89 GHz and of the MEMS off-state at 4 .. The radiation pattern with linear polarization was measured at $\Phi=0$ degree and $\Phi=90$ degree.

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## Chapter 4

## Beam-tilting EBG Dipole Base-station

This chapter describes EBG dipole array including the beam-tilting single element. This is realized with dipole antenna mounted on an EBG (Electromagnetic Bandgap) ground plane that can be changed by connecting several unit cells of the EBG. An array of four dipole-EBG elements is designed demonstrating a fixed down-tilt main beam for the base-station like antenna performance. The proposed array enables beam-tilting with a high directivity by utilizing the beam-tilting EBG-dipole single element. The beam-tilting of the EBG-dipole single element is investigated with a simple dipole and a printed dipole antenna. The high directivity more than 11 dB is measured at the angle of $30^{\circ}$ at 3.5 GHz. Higher directivity is achieved, compared to the reference case that uses the broadside EBG-dipole single element. In addition, the EBG-dipole array that can be mechanically adjusted is provided. The array antennas exhibit a reasonable impedance matching across the operational frequency.

## 4. 1 Introduction

The concept that offers the efficient beam scanning is suggested, and to verify the
concept, a new array design with dipole on the tunable EBG ground is presented. A beam-tilting single element is realized with a simple wire-dipole antenna and dipole mounted on the EBG that can be tuned by connecting several unit cells. The element enables to provide the fixed beam-tilt angle, which is suited for the base-station implementation. To validate the proposed array concept, the single element is extended to construct the four dipole-EBG array, including the progressive phase shift. For a fair comparison, the similar array incorporated with non-tuned standard EBG (single element radiates toward broadside) is designed as a reference case. A higher directivity at the desired angle and lower sidelobe are achieved, compared to the reference case. For the fixed scan angle or broadside, a reasonable gain from 11.2 dB to 11.4 dB is obtained, which is similar to or higher than the one similar four element array has. In addition, a dipole mounted on the standard EBG fed by equal phase feed is also provided for implementing the mechanically adjusted EBG-dipole array antenna. Both robust implementation and high performance within a wide angular range were achieved with the proposed dipole array antennas.

## 4. 2 Beam-tilting Wire-dipole on the EBG Structure

In this chapter, the beam-tilting wire-dipole antenna is investigated to verify the beamtilting ability as a single element for base-station applications. In the recent past, spatial beam-tilting array has attracted more attention in realizing an efficient base-station application for wireless communication. The radiation pattern can be tilted along the Eplane for enhancing the signal reception or along the H-plane for avoiding the co-channel interference between the adjacent base-station antennas [1-2]. To realize this spatial
beam-tilting array, the phased array has been effectively utilized by changing its array pattern [3]. This chapter describes a dipole antenna on miniaturized tunable EBG ground plane that can be used to reconfigure the radiation pattern along E-plane and H-plane. By tuning the EBG unit cell, the radiation pattern of the dipole antenna can be tilted toward a desired direction [4]. The designed antenna acts as a single element and this concept can be effectively extended for the base-station array antenna that has a fixed tilted beam after an installation. Using the proposed antenna, both increasing the signal reception and rejecting the adjacent beam can be realized.

## 4. 2. 1 Wire-dipole on Miniaturized EBG Structure

Mushroom-like structure has been widely used as a type of EBG ground planes [5]. In particular, the structure is useful to realize a low profile wire antenna placed on ground plane. The PMC-like EBG ground plane leads to the improved performance for the dipole which is parallel to the ground plane. Figure 4.1 (a) depicts the dipole antennas that are situated on the $6 \times 6$ EBG ground and PEC ground plane. $8 \times 8$ EBG ground plane given in [5] is miniaturized to construct $6 \times 6$ EBG-inspired ground plane, which has a bandgap around 4.25 GHz. Figure 4.1 (b) shows the S11 characteristics for dipole antenna situated along EBG and PEC grounds. It is seen that the EBG ground enhances the resonance of the parallel dipole, compared to that on PEC ground plane in spite of its miniaturized size. This miniaturized EBG can be tuned for tilting the radiation pattern.

(a)

(b)

Figure 4.1: Dipole antenna on EBG and PEC ground plane: (a) Configuration of dipole on EBG ground and PEC ground plane, (b) S11 characteristics for the dipole on different ground planes.

## 4. 2. 2 Beam-tilting Wire-dipole Antenna Design

The reconfigurable antennas allow the increasing demand for wireless communication to be met [6-7]. A tunable EBG ground can be applicable to the beam-tilting reconfigurable antenna. A tunable metal texture can tune the surface impedance of metal texture, resulting in the boundary condition between PMC and PEC surface. Ultimately, surface
wave properties due to the boundary condition can be controlled for tilting the beam [89]. Figure 4.2 (a) depicts a dipole antenna on the miniaturized tunable EBG-inspired ground plane. The dipole antenna is located at $0.05 \lambda$ above the $6 \times 6 \mathrm{EBG}$ ground on the 3.175 mm thick substrate with $\varepsilon_{\mathrm{r}}=2.2$. In this work, four standard EBG unit cells are combined to form a large patch element using an ideal metal strip. This large patch element allows the surface wave at the resonating frequency of the dipole while the surface wave on standard EBG is suppressed. Depending on where large patch elements are situated, the scan plane and the beam tilting direction can be determined. Figure 4.2 (b) and (c) show the EBG ground plane that includes the tuned large patch elements. The dimension of the EBG dipole follows as below:

$$
\begin{gathered}
\mathrm{L}=65.08 \mathrm{~mm}, \mathrm{H}=3.175 \mathrm{~mm} \mathrm{~W}_{\mathrm{S}}=9.53 \mathrm{~mm}, \mathrm{G}_{\mathrm{S}}=1.59 \mathrm{~mm}, \mathrm{D}_{\mathrm{S}}=0.79 \mathrm{~mm}, \mathrm{~W}_{\mathrm{L}}= \\
22.23 \mathrm{~mm}, \mathrm{G}_{\mathrm{L}}=4.76 \mathrm{~mm}
\end{gathered}
$$

The large patch elements are situated in parallel with the dipole or vertical to the dipole, from which radiation pattern is tilted along E-plane or H-plane, respectively. Figure 4.3 depicts the $\mathrm{E}_{\mathrm{y}}$ component on the surface of various EBG ground when dipole is placed along the $y$-axis and the simulated corresponding radiation pattern for the each scenario. Due to the miniaturized size of the EBG, some surface wave is observed above the EBG ground. For an original EBG ground plane, surface wave is shown to be evenly distributed, which allows a broadside radiation. For the tuned EBG ground planes, the surface wave at the side of tuned large elements is higher than the one at standard EBG unit cells. It allows the radiation pattern to be tilted toward the large elements. Using full-
wave simulator, HFSS, the tilting angle from $25^{\circ}$ to $29^{\circ}$ is achieved for the vertically and horizontally tuned EBG ground planes, respectively.


Figure 4.2: Topology of the dipole on tunable EBG ground plane: (a) Original $6 \times 6$ EBG ground, (b) Vertical tunable EBG ground, (c) Horizontal tunable EBG ground.


Figure 4.3: Surface wave for different EBG ground at 0.2 mm above the EBG ground and corresponding radiation pattern at 4.25 GHz .

## 4. 2. 3 Measurement results and discussion

The prototypes for five scenarios are depicted in Figure 4.4. These scenarios are determined, depending on where one dimensional periodic large patch elements are placed on the EBG ground plane.

The reflection coefficients and radiation pattern for the three prototypes are measured. Figure 4.5 shows the measured reflection coefficients for the each scenario. Although the resonance is shifted according to the different scenarios, the reflection coefficients $S_{11}$ is less than -10 dB at the center frequency of 4.25 GHz for all cases. For a convenience of the evaluation, the center frequency is chosen as 4.25 GHz , however, it is also possible to scale the antenna to be operated at other frequencies.


Figure 4.4: Prototypes of beam-scanning dipole on tunable EBG and scenarios for the measurement.


Figure 4.5: Measured Reflection coefficients for each scenario of the proposed antenna.

The antenna measured is mounted on the supporting structure as shown in Figure 4.6. The radiation pattern of the proposed antenna is measured and depicted in Figure 4.7. For the scenario I and II, the angles $\theta$ for the maximum radiation are $+25^{\circ}$ and $-25^{\circ}$ along the E-plane, respectively. An improvement of the radiation pattern at the angle $+25^{\circ}$ and $-25^{\circ}$ is around 1.0 dB , compared to the radiation pattern for the original EBG ground plane. For the scenario III and IV, a wider scan range can be attained as $+29^{\circ}$ and $-29^{\circ}$ along the H-plane, respectively. At the angle, $-29^{\circ}$ and $+29^{\circ}$, an enhancement around 2 dB over the original radiation pattern is achieved. The signal strength toward a certain direction can be improved by combining this element pattern with the corresponding array pattern. The main beam for the base-station application is typically fixed at $0^{\circ}-25^{\circ}$ for E-plane and scanned within $120^{\circ}$ for H-plane. The table for the gain of each scenario is provided in
the Table 4.1 The antenna gain ranging from 6.4 to 7.4 dB is attractive as a single element for the base-station applications that must achieve a narrow beam-width.


Figure 4.6: EBG dipole antenna mounted on the positioner for measurement.

Table 4.1: Antenna gain comparison

|  | Original EBG | Scenario I \& II | Scenario III \& IV |
| :---: | :---: | :---: | :---: |
| Maximum gain | 7.4 dBi | 6.4 dBi | 7.2 dBi |


(a) Dipole on the standard (original ) EBG

(b) Parallel tuned large elements to the dipole

(C)Vertical tuned large elements to the dipole

Figure 4.7: Measured radiation pattern of the proposed dipole antenna on the original EBG and the tuned EBG ground plane.

## 4. 3 Beam-tilting Printed Dipole on the EBG Structure

In this Chapter, the printed quasi-Yagi dipole antenna on the substrate is mounted on the tunable EBG structures. This design provides more robust implementation of the beamtilting EBG-dipole antenna, and can be extended to realize the beam-tilting array antennas for base-station applications.

## 4. 3. 1 Printed Dipole Antenna Mounted on EBG Structures.

The configuration of the EBG dipole antenna that acts as a single element is depicted in Figure 4.8. The half-wavelength dipole operated around 3.5 GHz is mounted on the different $6 \times 6$ miniaturized EBG structures: 1) EBG structure with standard unit cells and 2) tuned EBG structure that consists of standard EBG and the tuned EBG unit cells. For the modified EBG structure, $2 \times 6$ standard unit cells at the edge are replaced by the $1 \times 3$ tuned large element in order to change the radiation pattern. This tuned unit cell is realized by connecting four standard unit cells. The substrate used in the design is 1.27 mm and 1.9 mm thick RO6006 for antenna layer and EBG ground, respectively. The configuration of the dipole and EBG ground is shown in Figure 4.9, and a specification for the design is summarized in Table 4.2.


Figure 4.8: Printed dipole antenna mounted on a) standard EBG structure and b) tuned EBG structure.


Figure 4.9: Specification of the EBG-dipole element that consists of the dipole antenna and EBG structure.

Table 4.2: Antenna Dimension of the Single EBG-dipole

| $\mathrm{L}_{1}$ | 37.69 mm | $\mathrm{~L}_{5}$ | 1.66 mm | $\mathrm{H}_{4}$ | 35.68 mm | $\mathrm{G}_{2}$ | 1.31 mm |
| :--- | ---: | ---: | ---: | ---: | ---: | ---: | ---: |
| $\mathrm{~L}_{2}$ | 11.19 mm | $\mathrm{H}_{1}$ | 7.45 mm | $\mathrm{H}_{5}$ | 1.95 mm | $\mathrm{G}_{3}$ | 3.06 mm |
| $\mathrm{~L}_{3}$ | 4.37 mm | $\mathrm{H}_{2}$ | 4.43 mm | $\mathrm{~W}_{1}$ | 53.74 mm |  |  |
| $\mathrm{~L}_{4}$ | 2.52 mm | $\mathrm{H}_{3}$ | 18.35 mm | $\mathrm{G}_{1}$ | 0.53 mm |  |  |

## 4. 3. 2 Printed Dipole Antenna and Tunable EBG Structure

The dipole antenna is designed, based on the quasi-Yagi dipole antenna presented in [10] The current density on the surface of dipole is simulated to evaluate the dipole-like current distribution. Figure 4.10 shows the current distribution for the dipole without and with EBG ground plane, where similar magnitude of current on a pair of arms forms in the same direction. It indicates that the dipole antenna functions as a standard dipole antenna. The different configurations of the EBG ground largely determines the radiation characteristics of the dipole antenna. The dipole antenna on the standard EBG radiates toward the broadside while the dipole on the tuned EBG directs the radiation with a tilted angle from the broadside.


Figure 4.10: Simulated current density on the both arms of the dipole antenna, which is formed in the same direction (at 3.4 GHz for the dipole itself and at 3.5 GHz for the dipole mounted on EBG).

To offer an understanding on beam tilting mechanism, the bandgap properties near the resonant frequency of the structure is studied using full-wave simulation HFSS. The bandgap of the EBG structure can be determined using the reflection phase response across frequency bands [10-11]. Figure 4.11 shows the setup of the simulation and reflection phase response for the standard EBG and the tuned EBG unit cells $\left(\mathrm{G}_{3}=2 \mathrm{~mm}\right)$. For the standard EBG, the phase response around 3.5 GHz falls within the bandgap region that is defined in [11] while the tuned EBG shows the characteristics outside the bandgap. Due to the outside bandgap properties, the tuned EBG allows the surface wave which is tangential to the surface. This contributes to steer the radiation pattern from the broadside to the endfire. By replacing standard EBG unit cell with large EBG, therefore, it is possible to tilt the beam from the normal and toward where the tuned EBG is located.


Figure 4.11: Phase response versus the frequencies: the standard EBG unit cell located within the bandgap and the tuned EBG unit cell outside bandgap region.

## 4. 3. 3 Performance of the EBG-dipole Element

Next step is to study the performance of the complete EBG-dipole antenna. Figure 4.12 and Figure 4.13 shows the simulated impedance matching and total radiation pattern of the dipole antenna on the different EBG ground described in Figure 4.9. With respect to $7.5 \mathrm{~dB} \mathrm{~S}_{11}$, the printed dipole antenna covers the frequency bands: $3.34-3.5 \mathrm{GHz}$, and mounting of the dipole on different EBG structures shifts and broadens the bands to cover 3.43-3.8 GHz. For the broad-side radiation, a maximum gain of 4.7 dB is obtained with the standard EBG ground. For tuned EBG ground, It can be seen that the tuned EBG provides the gain of 4.9 dB at the desired angle of $\theta=30^{\circ}$. Note that the radiation patterns are simulated at 3.5 GHz where the maximum resonance is observed. This single element enables to offer a suitable beam-tilting for base-station applications.


Figure 4.12: S-parameters for dipole antenna, dipole on standard EBG, and dipole on tuned EBG.


Figure 4.13: Total radiation pattern on E- and H-plane: a) Dipole antenna itself, b) Dipole on the standard EBG sturucture, and c) Dipole on the tuned EBG.

## 4. 4 Beam-tilting EBG Dipole Array for the Base-station

The EBG dipole element presented in the Chapter 4.3 is extended for the EBG-dipole array antenna for the base-station applications. Based on the single element design, several base-station antennas are designed, and the performance of the array antennas is evaluated by using a simple array theories. The performance is compared to the fullwave simulation results.

## 4. 4. 1 EBG Dipole Array Design

Figure 4.14 depicts the configuration of the beam-tilting EBG dipole array antennas. The specification of the array antennas is presented in Table 4.3. The array antenna comprises four printed simple dipole antennas which are mounted on the EBG ground plane. The dimension of the array antenna is $(\mathrm{L} \times \mathrm{W} \times \mathrm{H})=219 \mathrm{~mm} \times 54 \mathrm{~mm} \times 72 \mathrm{~mm}$. This array antenna is based on the single element design discussed in Chapter 4.3. As discussed above, the tuned EBG ground is aimed at the down-tilt of the element pattern, and thus a better tilting performance of the array can be obtained. Another proposed design utilizes the standard EBG and equally phased array which can be mechanically tilted. Both dipole antenna and EBG ground are printed on 1.27 mm and 1.9 mm thick RO6006 substrate $\left(\varepsilon_{\mathrm{r}}=\right.$ 6.1), respectively. Those two substrates intersect through four holes made in the EBG structure where both PEC grounds on backside of substrates are electrically attached to form a common ground plane. Note that the thickness of the printed dipole is carefully selected in order to prevent loss of the EBG unit cells from the intersection.


Table 4.3: Antenna Dimension of the EBG-dipole array antenna

| $\mathrm{L}_{6}$ | 53.74 mm | $\mathrm{~L}_{10}$ | 67.75 mm | $\mathrm{H}_{8}$ | 15.53 mm | L | 218.74 mm |
| :--- | :--- | :--- | ---: | :--- | ---: | :--- | ---: |
| $\mathrm{~L}_{7}$ | 19.25 mm | $\mathrm{~L}_{11}$ | 202.75 mm | $\mathrm{H}_{9}$ | 9.80 mm | H | 72.36 mm |
| $\mathrm{~L}_{8}$ | 33.25 mm | $\mathrm{H}_{6}$ | 3.13 mm | $\mathrm{H}_{10}$ | 5.65 mm |  |  |
| $\mathrm{~L}_{9}$ | 39.75 mm | $\mathrm{H}_{7}$ | 2.43 mm | W | 53.74 mm |  |  |

The inter-spacing between elements is set to $0.623 \lambda$, which is approximately 54 mm , based on the size of the EBG-dipole element. The use of the $6 \times 6$ miniaturized EBG ground attributes to this inter-spacing restriction respective to the appearance of the grating lobe. Even though it may result in the degradation of the performance in impedance matching and radiation pattern, a huge gain loss caused by the grating lobe can be avoided.

## 4. 4. 2 Dipole Array Antenna Evaluation

The performance of the dipole array antenna itself without any EBG strucure is evaluated using full-wave simulation. In order to realize the efficient down-tilt of the main beam into a mobile device near the ground, the array antenna utilizes the down-tilt array pattern or mechanically tilted broadside array pattern, which has the maximum at the same point of the both element patterns. First step is to obtain a progressive phase difference $\alpha$ for each element. Next the phase shift attained can be applied to the feed network, adjusting the length of microstrip feed line. This feed network incorporated with the phase shift allows the array pattern to be tilted toward $30^{\circ}$, which is the tilting angle for the EBG-
dipole element. The equally phased feed network is also designed in order to achieve the mechanically tilting array antenna. For a better array design, it is desirable to evaluate the performance of the dipole array itself without EBG structure. The dipole arrays which includes 1) the progressive phase shift and 2) equal phase are simulated, and those feed network designs are depicted in Figure 4.15. Both feed network follows the dimension presented in Table 3, except for the difference in a symmetric configuration of the microstrip line and a slight tuning of the equally phased array in order to realize a desired impedance matching characteristics of the dipole array antenna. The specification of the equally phased array is also shown in Figure 4.15. The modified dimension follows as below:

$$
\mathrm{L}_{12}=23.5 \mathrm{~mm}, \mathrm{~L}_{13}=1.86 \mathrm{~mm}, \mathrm{~L}_{14}=3.5 \mathrm{~mm}, \mathrm{G}_{4}=0.3 \mathrm{~mm}
$$

In Figure 4.16, the impedance matching of the array is presented, and both phase shifted and equally phased arrays achieves $10 \%$ and $5 \%$ bandwidth, respectively. With the phase shifted array case, it can be observed that a better impedance matching is achieved than the equally phased array as discussed in [12]. The total radiation pattern for the dipole array is shown in Fig. 11. The maximum of the radiation pattern directs into a broadside with equally phased array while the phased array a tilted angle of $30^{\circ}$. The gain of those two arrays are similar as around 5 dB , since element pattern of the dipole antenna is more omni-directional than EBG-dipole antenna, which results in less gain loss to tilt the radiation pattern.


Figure 4.15: Configuration of the dipole array (no EBG structure) incorporated with a) an equal phase feed network and b) a progressive phase feed network.


Figure 4.16: S-parameter characteristics for dipole array antenna without EBG.


Figure 4.17: Simulated total radiation pattern in the H-plane (main beam-tilting plane).

## 4. 4. 3 Array Analysis and Full-wave Simulation Results

Simple array analysis is performed based on the single element pattern provided the Chapter 4.3. The full EBG-dipole array antenna is simulated using full-wave simulation, HFSS. The array antenna consists of the four dipole array antennas and EBG structure which are shown in Figure 4.15 and Figure 4.18, respectively. There are four EBG-dipole antennas that can be designed based on different combinations of dipole array and EBG structure. Among those four cases, three meaningful cases including two proposed antenna designs are examined.


Figure 4.18: Configurations of the EBG structure: a) standard EBG and b) tuned EBG structure.

The examined three meaningful designs are presented in Figure 4.18. The three cases studies are listed as below:

1) Standard EBG structure and progressive phase (Reference)
2) Tuned EBG structure and progressive phase feed
(Proposed antenna design I: without mechanical adjustment)
3) Standard EBG structure and equal phase feed
(Proposed antenna design II: with mechanical adjustment)


Figure 4.19: Case studies through the full-wave simulation: a) Standard EBG and progressive phase shift (reference: conventional array antenna) b) Tuned EBG and progressive phase shift (proposed design I), c) Standard EBG with equal phase (proposed design II: mechanically adjusted array antenna).

Those cases are simulated through using full-wave simulation. Figure 4.20 shows the impedance matching of the three cases. The frequency bands: $3.35-3.85 \mathrm{GHz}\left(\mathrm{S}_{11}<7.5\right.$ dB ) is covered by the tuned EBG and progressive phase case, which will be compared to the measured results later. For the purpose of the evaluation, the similar four element array pattern, based on the array analysis, is calculated. By invoking progressive phase difference and element spacing discussed previously, the normalized array pattern can be defined as [13]

$$
\begin{align*}
& A F_{N}=\frac{\sin (N \Psi / 2)}{N \sin (\Psi / 2)}  \tag{4.1}\\
& \Psi=\beta d \cos \theta+\alpha
\end{align*}
$$

where $N=4$, inter-spacing $d=0.623 \lambda$, and $\alpha=\left\{\begin{array}{l}0 \text { (Broadside) } \\ 0.6 \pi \text { (Tilted case) }\end{array}\right.$
The main difference in the array analysis is to employ the single element pattern (simulated) which has the maximum toward 1) broadside or 2) tilted direction where array pattern points out, hence, the effect of the different element pattern on the total radiation can be studied. Figure 4.21 (a), (b), (c) shows the comparison between simulated results and the array analysis using the broadside and the tilted element pattern. Although it is not rigorous analysis, similar characteristics in the important parameters such as beam-width and first null position is observed. In Figure 4.20 (d), the radiation pattern for the different cases simulated at 3.5 GHz is compared. The maximum directivity of 11.2 dB is attained with the tuned EBG and progressive phase (with the tilted element: proposed design I), which is 1.5 dB higher than the standard EBG and progressive phase (with broadside element: reference). Considering that both tilted and
broadside element shows the similar directivities, it can be mentioned that proposed array concept leads to an improved directivity, compared to the conventional array antennas. The directivity achieved with the proposed array is only 0.2 dB less than the case of the standard EBG with equal phase feed. In following section, the link-budget evaluation of the EBG-dipole array antenna will be presented.


Figure 4.20: Simulated S-parameter characteristics for representative full dipole-EBG array antennas.


Figure 4.21: Comparison between simple array analysis and normalized simulated radiation pattern, and directivities at 3.5 GHz .

## 4. 5 Linkbudget Evaluation of the EBG Dipole Array

Radio blind areas are the location where the waves are rarely received. There are two kinds of radio blind areas: 1) closed areas and 2) open area. In a closed area, the waves are remarkably attenuated because of the blocking effect caused by the tunnel, inside building, and so on. The open blind areas occurs in the line-of-sight scenario when mobile device falls outside the edge area of the base-station. In this dissertation, linkbudget between base-station and mobile antennas in the line of sight is evaluated with the EBG-dipole array antennas presented in Chapter 4.

Near the edge area, the coverage range is greatly affected by the small change of the power transmission. In order to realize a communication link, the power transmission in between must be at least -90 dB [12]. In this range, an improved power transmission of 12 dB leads to the increase by $10 \%-20 \%$ of maximum coverage range. To verify the enhancement in the linkbudget, a comprehensive linkbudget evaluation is conducted with respect to the impedance matching and directivity of the array obtained at several operating frequencies. The simulated results provided in Chapter 4 is utilized in this evaluation. The communication link between base-station antenna and mobile phone antenna is depicted in Figure 4.22.


Figure 4.22: Description of the linkbudget analysis between base-station and mobile phone ( $\theta_{\mathrm{t}}=30^{\circ}, \theta_{\mathrm{r}}=60^{\circ}$, and R varies from 4 m to 200 m ).

The standard Friis power transmission formula [13] is used in this evaluation, which can be written as

$$
\begin{align*}
& S_{21}\left(R, \theta_{r}, \phi_{r}\right)=\frac{P_{r}}{P_{t}}=\frac{\lambda^{2}}{16 \pi^{2} R^{2}} G_{t}\left(\theta_{t}, \phi_{t}\right) G_{r}\left(\theta_{r}, \phi_{r}\right)  \tag{4.2}\\
& \times\left(1-\left|\Gamma_{t}\right|^{2}\right)\left(1-\left|\Gamma_{r}\right|^{2}\right)\left|\hat{\rho}_{t} \cdot \hat{\rho}_{r}\right|^{2}
\end{align*}
$$

where $G(\theta, \phi)$ is the far-field gain pattern with $(\theta, \phi)$ angular dependence, $1-|\Gamma|^{2}$ is the input impedance mismatch, and $\left|\hat{\rho}_{t} \cdot \hat{\rho}_{r}\right|^{2}$ is the polarization mismatch.

The linkbudget evaluation is conducted at three selected frequencies: $3.45 \mathrm{GHz}, 3.5 \mathrm{GHz}$, and 3.55 GHz . The impedance matching at those frequencies is already presented in Figure 4.20 , and the loss by the mismatch is considered for all cases. The simulated radiation pattern at 3.45 GHz and 3.55 GHz is shown in Figure 4.23. For a fair evaluation, a mechanical tilting adjustment toward $30^{\circ}$ is applied to the case of the standard EBG with equal phase. The maximum gains obtained around the down-tilt angle of $30^{\circ}$ are used in this evaluation, and it is assumed that there is no polarization mismatch between base-station and mobile antennas. The mobile antenna is assumed to utilize a simple dipole like radiation pattern. The radiation pattern can be approximated by

$$
\begin{equation*}
G_{r}\left(\theta_{r}, \phi_{r}\right)=G_{\text {max }, \text { dipole }} \times\left|F_{r}\left(\theta_{r}, \phi_{r}\right)\right|^{2}=1.5 \times\left|\sin \theta_{r}\right|^{2} \tag{4.3}
\end{equation*}
$$

where the dipole antenna points out the base-station at the angle of $\theta_{\mathrm{r}}=60^{\circ}$. The parameters incorporated into the Friis formula are summarized in Table 4.4.

Table 4.4: Parameters for the linkbudget evaluation

|  |  | 3.45 GHz | 3.50 GHz | 3.55 GHz |
| :---: | :---: | ---: | ---: | :---: |
| Standard EBG with | $\mathrm{S}_{11}$ | -9.5 dB | -11.6 dB | -19.0 dB |
| progressive phase | Gain | 9.0 dB | 9.7 dB | 10.2 dB |
| Tuned EBG with | $\mathrm{S}_{11}$ | -12.5 dB | -12.2 dB | -12.3 dB |
| progressive phase | Gain | 11.3 dB | 11.2 dB | 11.2 dB |
| Standard EBG with | $\mathrm{S}_{11}$ | -7.2 dB | -15.0 dB | -7.0 dB |
| equal phase | Gain | 8.8 dB | 11.4 dB | 12.2 dB |



Figure 4.23: The radiation pattern simulated at (a) 3.45 GHz and (b) 3.55 GHz that will be incorporated into the Friis power transmission formula.

The calculated communication linkbudget at multiple frequencies is presented in Figure 4.24, Figure 4.25, and Figure 4.26. It can be clearly seen that the tuned EBG with progressive phase always outperform the standard tuned EBG with progressive phase (reference: a traditional array) by $2.5 \mathrm{~dB}, 1.5 \mathrm{~dB}$, and 1.0 dB at $3.5 \mathrm{GHz}, 3.45 \mathrm{GHz}$ and 3.54 GHz, respectively.

For the case of the standard EBG with equal phase, an enhancement observed, compared to the reference case, is 1.7 dB , and 1.8 dB at 3.5 GHz , and 3.55 GHz , respectively. At 3.45 GHz , however, the coverage range is greatly reduced since the EBG-dipole incorporated with an equal phase has a narrow bandwidth characteristics which does not cover lower frequency band. The dimension of the EBG structure is originally optimized for the EBG-dipole array with progressive phase, however, the EBG structure is not optimized for the array with an equal phase. The solution to increase the bandwidth is to use bigger EBG ground plane which resonates at lower frequency, then, the equal phase feed case can also cover lower frequency bands. With respect to the effective coverage area, more remarkable improvement can be observed by utilizing the proposed arrays.

Let us consider that the minimum communication link is -75 dB to maintain a safe communication link. The coverage radius centered at the base-station can be estimated, based on the communication link. The minimum communication link is indicated in the Figure 4.24, Figure 4.25, and Figure 4.26, from which the effective range can be estimated. For instance, in the Figure 4.25, the radius of the effective coverage is increased from 120m (Standard EBG with progressive phase) to 143 m (Tuned EBG with
progressive phase) or to 147 m (Standard EBG with equal phase) at 3.5 GHz by employing the proposed antenna designs. It is implied that near the cell edge, an additional 1-2 dB in the communication link remarkably improves the coverage of the base-station. The radius of base-station coverage for all other cases is presented in Table 4.5. Some additional coverage can be obtained using the proposed array at different frequencies.


Figure 4.24: Power transmission level between representative dipole-EBG array antenna and mobile antenna at 3.45 GHz .


Figure 4.25: Power transmission level between representative dipole-EBG array antenna
and mobile antenna at 3.50 GHz .


Figure 4.26: Power transmission level between representative dipole-EBG array antenna and mobile antenna at 3.55 GHz .

Table 4.5: Radius of the Effective Coverage Area (Minimum Linkbudget= -75 dB )

|  | 3.45 GHz | 3.50 GHz | 3.55 GHz |
| :---: | :---: | :---: | :---: |
| Standard EBG with progressive phase | 118 m | 120 m | 127 m |
| Tuned EBG with progressive phase | 144 m | 143 m | 143 m |
| Standard EBG with equal phase | 80 m | 147 m | 151 m |

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## Chapter 5

## Circular Polarizer, Septum for HPM

This chapter describes the circular polarizer, septum which is profiled with smooth contour. The typical septum polarizer which has gained popularity in the literature may be unsuitable for high power applications due to the sharp corners in the design. In order to address this issue, the fundamentals of the septum operation are first revisited using a graphical visualization through full-wave analysis. Next, a septum profiled with smooth edges is presented with enhanced power handling capabilities in comparison to the stepped septum polarizer. In this work, the sigmoid function is introduced to represent the smooth contour of the septum and to enable diverse configurations without any discontinuities. The smooth and stepped profiles are optimized using the Particle Swarm Optimization (PSO) technique. The maximum electric field intensity around the smooth edges is investigated using a full-wave simulator, HFSS, and our observations show that the maximum electric field is reduced by $42 \%$ in comparison to the stepped septum. The numerical approach is evaluated by comparing the exact series solution for the half-plane scattering problem with the simulated results in HFSS. A septum design with rounded edges is also studied as another possible design to reduce the maximum fields. Communication systems using circular polarization are reported [1-2]. With practical
high power microwave sources coming closer to realization, it will be important to provide RF devices that can handle high power, and therefore this dissertation will focus on achieving better power handling in the septum design.


Figure 5.1: Circular polarization ground station reflector and septum polarizer profiled with smooth edges for possible use in the HPM system.

### 5.1 An Overview of the Sigmoid Septum for Handling HPM

In this dissertation, the septum is revisited with respect to handling high power microwave (HPM) [3-8] signals by using the sigmoid function [9] to smoothen the sharp edges. The sigmoid function was chosen because the function and its derivatives are continuous. This ensures that the curvature is continuous, which is important to decrease the electric fields in the surrounding area. As depicted in Figure 5.1, the proposed sigmoid septum polarizer that will be detailed in later chapter can be incorporated into the feed horn given in [10-11] as a possible feed for a large reflector antenna system. In our investigation, we optimize the sigmoid septum as well as the original stepped septum design in an effort to compare the two designs. One primary question in our research is whether a smooth septum can indeed achieve a similar performance, if not better, in comparison to the original stepped septum design. Therefore, both of these designs are optimized using the nature-inspired global optimizer, Particle Swarm Optimization (PSO) applied to antenna designs [12-15]. Since performance can depend on bandwidth, different frequency ranges were used for the fitness function, and the following terminology is used to describe each design investigated:

1) $25 \%$ sigmoid septum: sigmoid septum optimized within $25 \%$ bandwidth
2) $25 \%$ stepped septum: stepped septum optimized within $25 \%$ bandwidth
3) $5 \%$ sigmoid septum: sigmoid septum optimized within $5 \%$ bandwidth

Next, we validate the sigmoid septum by investigating the electric field intensity around the edges of the septum. In particular, the increased electric field intensity around the sharp edge can result in air breakdown inside the horn antenna. Therefore, a comparison
between the maximum electric fields observed near the smooth and stepped septums is made to evaluate the relative improvement by using a smooth septum. Significant reduction in electric field is observed for the new sigmoid septum in comparison to the stepped septum. Based on the reduced electric fields, the improvement of power capacity from the sigmoid septum is obtained. Each component in electric field is acquired at the position where total maximum electric field is induced, and the contribution to the maximum total electric field is discussed. To validate our numerical approach for obtaining the field data, the canonical half-plane scattering problem in HFSS is compared with the exact series solution for half-plane scattering. In addition, the stepped septum that utilizes rounded edges is investigated in terms of its power capabilities, and the electric field intensity on the septum plane is compared to the other septum designs.

## 5. 2 Numerical Evaluation for the Septum Polarizer

In this chapter, the fundamental operation of the septum polarizer is revisited using HFSS, a full-wave simulator. Port 1 and port 2 excitations can be decomposed into even and odd mode excitations, and the septum responses to even and odd mode excitations are graphically visualized [16-17]. The distribution of a circularly polarized wave is illustrated as it propagates through the septum polarizer. Figure 5.2 depicts the circular horn with the septum polarizer, which is located in the middle of the circular waveguide. For the circular waveguide, the septum polarizer divides the waveguide into two semicircular waveguides [18-25]. The two septum designs shown in Figure 5.2 have been optimized for on-axis axial ratio and S11.


Figure 5.2: Generation of the RHCP and LHCP waves depending on excitations at port 1 or port 2 for the sigmoid and stepped septum.


Figure 5.3: Decomposition into the even and odd modes for each port excitation.


Figure 5.4: Electric field distribution for the even and odd mode: the vertical electric field and horizontal electric field.

The circular polarization due to excitation at port 1 or port 2 is investigated by plotting the electric field vectors within the waveguide. Figure 5.2 shows the successful achievement of LHCP and RHCP waves using the sigmoid (smooth) and stepped septums. If exciting port 1 , an RHCP wave can be successfully achieved, while an LHCP wave is generated if port 2 is excited. Figure 5.3 shows the quasi-TE11 excitation at port 1 or port 2 which consists of two modes: an even mode and an odd mode. For the even mode, the directions of the two excitations are the same, while the excitations for the odd mode are reversed. By summing even and odd modes, an excitation either to the port 1 or port 2 can be determined.

The electric field responses to both even and odd modes are simulated, and the electric field distribution on the plane above the septum is depicted in Figure 5.4. When the even mode is excited, the two vertical electric fields propagate through the circular horn antenna, and the electric field distributions remain unchanged after the septum region. On the other hand, exciting the odd mode creates the horizontal electric field. By selecting one port of the circular horn antenna and exciting the port, both vertical and horizontal components can be obtained. A well-optimized septum polarizer can realize a $90^{\circ}$ phase difference and an equal magnitude between the two orthogonal electric fields. In order to achieve a good circularity, the phase difference between the two orthogonal components must be

$$
\begin{equation*}
\varphi_{z}-\varphi_{x}= \pm 90^{\circ}(\text { Phase difference }) \tag{5.1}
\end{equation*}
$$

In this dissertation, the electric field response (both phase and magnitude) on the septum plane is investigated as the wave propagates through the circular horn antenna. First, the


Figure 5.5: Generation of the circular polarization: visualization of the phase difference between $E_{X}$ and $E_{Z}$ components on septum plane.
phase difference between $E_{X}$ (horizontal) and $E_{Z}$ (vertical) components is studied along the septum plane. Figure 5.5 depicts the visualization of the phase difference for the stepped and sigmoid septums when port 1 is excited. It can be seen that $90^{\circ}$ phase difference between the $\mathrm{E}_{\mathrm{X}}$ and $\mathrm{E}_{\mathrm{Z}}$ components is achieved after the wave propagates through the septum region. This is because the vertical and horizontal modes have different propagation constants $\beta$ within the septum transition regions, and therefore one


Figure 5.6: Generation of the circular polarization: relative magnitude comparison between the $\mathrm{E}_{\mathrm{X}}$ and $\mathrm{E}_{\mathrm{Z}}$ components on the septum plane.
mode is delayed with respect to the other mode. Therefore, if the septum is well optimized, the $\pm 90^{\circ}$ phase difference between two orthogonal components can be achieved. Next, Figure 5.6 shows the variation of the magnitude along the sigmoid and stepped septum plane. It can be seen that the $\mathrm{E}_{\mathrm{X}}$ component builds up until it is equivalent to the $\mathrm{E}_{\mathrm{Z}}$ electric field component. After the septum region, the magnitude of the horizontal component is almost identical to the vertical component. With equal
magnitudes and $\pm 90^{\circ}$ phase differences, circular polarization is obtained. Therefore, it has been shown that a septum polarizer can properly modify the phase and magnitude of the two orthogonal components to generate a good circular polarization. In particular, this statement is valid when a well-designed septum polarizer is used. In the next section, the design of the septum polarizer using a global optimization technique will be discussed.

### 5.3 Optimizing the Smooth (Sigmoid) and Stepped Septum Using PSO

In this chapter, the septum polarizers having smooth and stepped edges are optimized using Particle Swarm Optimization (PSO). The PSO technique is implemented to optimize the geometry set of the septum polarizers that are loaded in a probe-fed circular horn antenna, which are scaled versions of the designs in [10]. The configuration of the horn is shown in Figure 5.7. Once PSO is applied, the configurations and geometry set of the optimized septum polarizers are attained.

### 5.3.1 Sigmoid Function for the Smooth Edge of the Septum Polarizer

As discussed in Chapter 5.1, the stepped septum polarizer may not be suitable for a HPM system. The smooth edge of the septum is more desirable to reduce the electric field intensity near the edge and avoid air breakdown inside a horn antenna.


Figure 5.7: The circular horn antenna loaded with the septum polarizer for Particle Swarm Optimization (PSO) run.

The sigmoid function is used in this paper to smooth the septum contour. One expression of the sigmoid function can be written as

$$
\begin{equation*}
S(x)=\frac{H}{1+e^{4(L-x) / H C}} \tag{5.2}
\end{equation*}
$$

where $L, H$, and $1 / C$ represents the position at center, height, and slope of the sigmoid, respectively. The sigmoid function is advantageous to construct a septum polarizer for the HPM system due to its unique features such that:

1) It can provide diverse profiles changing from a step to smooth curves through the variable C which determines the slope of each step.
2) A smooth profile without any discontinuities in curvature can be achieved using a relatively small number of variables, lowering the dimensionality of the optimization problem.
3) For our optimization problem, the goal is to provide a function which fits within the specified waveguide dimensions. The sigmoid function rapidly approaches its final values, and can easily fill the waveguide without any later adjustment.

To the authors' best knowledge, there has not been any study published in literature investigating a smooth septum. Therefore one of our goals in this research is to determine whether the sharp edges are required for a good septum performance. As mentioned in the first item in the list, the sigmoid function is able to achieve the stepped contour by setting $\mathrm{C}=0$. This allows the optimizer to choose the best design without prior bias. For example, if the final optimized design does not have sharp corners then one might conclude that sharp corners are not necessary for good performance. The second item in the list is important to enhance the power handling capabilities of the septum. We could have easily chosen to use rounded corners or even Piecewise Monotonic Cubic Polynomials, but the second derivative would then be discontinuous, leading to higher electric fields. Therefore we chose the sigmoid function because of its continuous zeroth, first, and second derivatives. The last item in the list addresses the need for the function to fit within a specified height H . Many other functions exist that are sigmoidal in shape but decay much slower than the sigmoid function. This is useful because it allows us to specify the total sigmoid height by the variable H , rather than using the function value $S(\mathrm{x})$ at the points of interest. Overall, the sigmoid function satisfies many of the criterion needed to simplify the optimization and reduce the electric field strength.

Figure 5.8 (a) depicts some examples of sigmoid functions with different positions,


Figure 5.8: a) Diverse profile of the sigmoid functions: from the stepped to the smooth profile. The black curves demonstrate the effect of the slope C , the blue curves show the effect of the height H , and the green curves demonstrate the effect of the location L. b)

Taylor series expansion of the sigmoid function near the point $x=L$.


Figure 5.9: Illustration of the 6 sigmoid function and its comparison to a possible 6 stepped function.
heights, and slopes of each step. One useful property of the representation in equation (5.2) can be seen from the Taylor series expansion of $\mathrm{S}(\mathrm{x})$ at the point $x=L$, which can be written as

$$
\begin{equation*}
S(x) \approx \frac{H}{2}+\frac{1}{C}(x-L) \tag{5.3}
\end{equation*}
$$

Clearly, the $C$ parameter controls the slope of the sigmoid, as shown in Figure 5.8 (a) and (b). One can also observe the $H$ and $L$ parameters control the height and position of the sigmoid without changing the slope. This allows an easier formulation of the optimization boundaries used within PSO. With the sigmoid function, the profile of the function can be changed from a stepped to smooth curves. In particular, one form of the
sigmoid function can be a stepped function by setting $C=0$. The sigmoid function in equation (5.3) only represents one sigmoid step, but more sigmoids must be incorporated into the function in order to achieve a multi-sigmoid step function. One can accomplish this through a summation of sigmoids. Each sigmoid function with a different position, slope, and amplitude can be combined to represent the contour of the septum as written below as

$$
\begin{equation*}
S_{\text {combined }}(x)=\sum_{i=1}^{6} \frac{H_{i}}{1+e^{4\left(L_{i}-x\right) / H_{i} C_{i}}} \tag{5.4}
\end{equation*}
$$

By combining the sigmoid functions in this fashion, the continuity of the zeroth, first, and second derivatives is preserved throughout the region of interest. The sigmoid function consists of six steps, and Figure 5.9 shows an example of the implementation of the stepped and sigmoid contour of the septum using equation (5.4). The septum polarizer profiled with this function will be optimized using the PSO technique.

## 5. 3. 2 Particle Swarm Optimization (PSO) Applied to the Septum Polarizer

Particle Swarm Optimization (PSO) has demonstrated a wide applicability for providing final design solutions for electromagnetic devices such as corrugated horns [12], highperformance array antennas [13], and wireless communication antennas [14]. In each of these applications, PSO was successfully implemented to achieve enhanced antenna performance. Its success can be attributed to its ability to solve nonlinear, ill-conditioned, multimodal, discontinuous, non-differentiable, non-convex, and highly multidimensional problems, which are typical in electromagnetics [15]. Our goal in this research is to apply the PSO technique to a stepped septum and a sigmoid septum to compare their optimized
performance. In this optimization process, our primary goal is to achieve a good circularity and impedance matching of the septum polarizer. The fitness function bridges the PSO optimizer and the full wave simulator, HFSS, and describes the antenna performance for a given geometry. In this work, the fitness function is configured to lead to the optimum design that has a minimum VSWR and axial ratio (AR) within a particular bandwidth. The fitness function can be expressed as

$$
\begin{equation*}
f(\vec{H}, \vec{C}, \vec{L})=\underset{f_{i} \in F}{\operatorname{Max}}\left(\left.V S W R\right|_{f_{i}}\right)+\sqrt{2} \operatorname{Max}_{f_{i} \in F}\left(\left.A R\right|_{f_{i}}\right) \tag{5.5}
\end{equation*}
$$

where $\vec{H}, \vec{C}$, and $\vec{L}$ are vectors that consists of the geometrical parameters of the six variables. The $\sqrt{ } 2$ factor is used so that a similar emphasis is placed on both the AR and the VSWR. Note that the maximum electric field is not being included due to its heavy computational burden, using the procedure as outlined previously. However, better power capacity will still be achieved assuming that none of the optimal values for C are zero. The septum polarizer with the circular horn antenna is simulated in HFSS for the optimization. To excite each semi-circular waveguide, two coaxial feed ports are located at both sides of the semi-circular waveguide as shown in Figure 5.7. The simulated VSWR and AR for numerous possible geometries are examined by PSO. Ultimately, implementing PSO allows us to find geometries that can minimize the fitness value presented in (5.5). Extensive computing resources are typically required to evaluate the fitness function. For PSO, one can easily accelerate the evaluation process by using a parallel optimization process.

The fitness function is established to optimize the septum polarizer within a particular bandwidth which is centered at 5.8 GHz . The following frequency ranges are used for the fitness function in the optimization:

1) Sigmoid septum optimized within $5.22 \mathrm{GHz}<f_{i}<6.38 \mathrm{GHz}$ ( $25 \%$ bandwidth).
2) Stepped septum optimized within $5.22 \mathrm{GHz}<f_{i}<6.38 \mathrm{GHz}$ ( $25 \%$ bandwidth).
3) Sigmoid septum optimized within $5.66 \mathrm{GHz}<f_{i}<5.95 \mathrm{GHz}$ ( $5 \%$ bandwidth).

The work in $[19,21]$ reports that the stepped septum polarizer can achieve a $25 \%$ bandwidth with a good performance. A similar bandwidth is applied to the both sigmoid and stepped septums, and a narrower bandwidth is considered for another sigmoid septum in order to investigate its performance for narrower bandwidths.

## 5. 3. 3 PSO-Optimized Sigmoid and Stepped Septum Polarizers

As mentioned above, the sigmoid and stepped septum polarizers with six steps are optimized. This optimization is aimed at achieving a good axial ratio and impedance matching within a certain bandwidth. Before this investigation, it was unknown whether the smooth septum could achieve a similar bandwidth to the stepped septum. To observe the influence of the bandwidth setup on the septum profile, different bandwidths (5\% and $25 \%)$ at 5.8 GHz are established for the sigmoid septum polarizer. Therefore, one stepped and two sigmoid septum polarizers, namely $25 \%$ stepped septum, $25 \%$ sigmoid septum, and 5\% sigmoid septum are optimized using PSO. The PSO program also monitored the change in fitness during its run, and the optimization was terminated once stagnation was
observed for a given number of runs. After 350 iterations, the fitness value for all three septum polarizers show minute changes in fitness, since the PSO run is in the local optimization phase. Therefore, the optimum configuration of the septum can be found around 350 iterations.

The dimension of the final optimum septum polarizers is provided in Table 5.1. For the sigmoid septum, height $H$, length $L$ and variable $C$ of each step are provided. However, the variable C is set to be zero for the stepped septum. The configuration of the optimized stepped and sigmoid septums are depicted in Figure 5.10. It can be observed that the sigmoid septum requires roughly $0.1 \lambda$ longer septum length than the stepped septum to achieve the $90^{\circ}$ phase difference between the two orthogonal components. In addition, a difference can be observed between the two profiles of the sigmoid septums in terms of the smoothness of the profile on average. The variables C for the $5 \%$ sigmoid design are slightly larger than the variables C for the $25 \%$ sigmoid septum design, which indicates that the $5 \%$ sigmoid has a slightly smoother profile than $25 \%$ sigmoid septum. From this, one might conjecture that PSO provides a smoother profile of the sigmoid septum if a narrower bandwidth is incorporated in the fitness function, which may imply that there is a tradeoff between bandwidth and power handling capacity. The smoother profile of the $5 \%$ sigmoid than $25 \%$ sigmoid can be understood from the above discussion. In the following sections, the optimized sigmoid and stepped septum polarizers above will be used in the evaluation.

Table 5.1 Optimized Geometry Set for Sigmoid and Stepped Septum Polarizer

| $\mathrm{V}_{1}$ | $2.4581 \lambda$ | $\mathrm{~W}_{2}$ | $1.3167 \lambda$ |
| :---: | :---: | :---: | :---: |
| $\mathrm{~V}_{2}$ | $0.5223 \lambda$ | $\mathrm{~W}_{\mathrm{P}}$ | $0.2665 \lambda$ |
| $\mathrm{~V}_{3}$ | $1.3167 \lambda$ | T | $0.0614 \lambda$ |
| $\mathrm{~W}_{1}$ | $0.7949 \lambda$ | Septum <br> thickness | $0.03 \lambda$ |
|  | SIGMOID | SIGMOID |  |
|  | SEPTUM WITH | SEPTUM WITH | STEPPED SEPTUM |
|  | $5 \%$ BW | $25 \%$ BW | WITH 25\% BW |
| $\mathrm{L}_{1}$ | $0.3159 \lambda$ | $0.2029 \lambda$ | $0.3467 \lambda$ |
| $\mathrm{~L}_{2}$ | $0.7349 \lambda$ | $0.6073 \lambda$ | $0.7079 \lambda$ |
| $\mathrm{~L}_{3}$ | $0.9004 \lambda$ | $0.9270 \lambda$ | $0.9997 \lambda$ |
| $\mathrm{~L}_{4}$ | $1.1734 \lambda$ | $1.1668 \lambda$ | $1.1961 \lambda$ |
| $\mathrm{~L}_{5}$ | $1.4482 \lambda$ | $1.4313 \lambda$ | $1.3770 \lambda$ |
| $\mathrm{~L}_{6}$ | $1.6574 \lambda$ | $1.5339 \lambda$ | $1.4681 \lambda$ |
| $\mathrm{H}_{1}$ | $0.1556 \lambda$ | $0.1225 \lambda$ | $0.1071 \lambda$ |
| $\mathrm{H}_{2}$ | $0.0401 \lambda$ | $0.1044 \lambda$ | $0.1126 \lambda$ |
| $\mathrm{H}_{3}$ | $0.0295 \lambda$ | $0.0919 \lambda$ | $0.1182 \lambda$ |
| $\mathrm{H}_{4}$ | $0.1851 \lambda$ | $0.0646 \lambda$ | $0.0626 \lambda$ |
| $\mathrm{H}_{5}$ | $0.1552 \lambda$ | $0.0259 \lambda$ | $0.1981 \lambda$ |
| $\mathrm{H}_{6}$ | $0.2484 \lambda$ | $0.4046 \lambda$ | $0.1963 \lambda$ |
| $\mathrm{C}_{1}$ | 0.4211 | 0.1057 | 0 |
| $\mathrm{C}_{2}$ | 0.7496 | 0.5161 | 0 |
| $\mathrm{C}_{3}$ | 0.9405 | 1.1171 | 0 |
| $\mathrm{C}_{4}$ | 0.4234 | 0.4384 | 0 |
| $\mathrm{C}_{5}$ | 0.9339 | 1.0658 | 0 |
| $\mathrm{C}_{6}$ | 0.8294 | 0.4767 | 0 |




## Optimized 6 stepped septum for $25 \%$ bandwidth

Figure 5.10: Final optimized configurations (scaled) of the sigmoid septum optimized for bandwidths of $25 \%$ and $5 \%$, and the stepped septum optimized for a the bandwidth of $25 \%$.

### 5.4 Performance of the Septum Polarizer

The C-band horns loaded with the optimized sigmoid and stepped septums are evaluated using HFSS, a full-wave simulator. The evaluation is performed over the bands of interest which have a center frequency at 5.8 GHz . The on-axis axial ratio, radiation pattern, reflection coefficient, and isolation between the two coaxial feed ports are simulated, and the influence of the smooth profile on the broad bandwidth performance is studied.

## 5. 4.1 Septum Polarizer Performance

Figure 5.11 shows the on-axis axial ratio for the optimized sigmoid and stepped septums across a 1.5 GHz bandwidth (roughly $25 \%$ at 5.8 GHz ).


Figure 5.11: Simulated on-axis axial ratio for sigmoid and stepped septum polarizer versus frequencies.

The circularity of the horn is dependent on the septum profile, where the septum geometry dictates the propagation constants for the two orthogonal modes. As discussed previously, one must ensure that one mode is delayed such that it is $90^{\circ}$ out of phase of the other mode. The bandwidth configuration in the PSO fitness function will also characterize how the septum profile is chosen. Optimizing over a wider bandwidth forces the PSO engine to search for designs that have good impedance matching and axial ratio over a wider bandwidth. This may come at a cost, where one must compromise between good AR/S11 performance versus bandwidth. We observe this in the final performance evaluation of each of these horns, which is plotted in Figures 5.11-5.16. We observe from these plots that the sigmoid design optimized for a smaller bandwidth ( $5 \%$ sigmoid) in general shows better performance over the band. However, the other designs provide a wider bandwidth in both S 11 and AR. For a 1 dB axial ratio criterion, the $25 \%$ sigmoid septum, $25 \%$ stepped septum, and $5 \%$ sigmoid septum provide a good circularity across $25 \%, 15 \%$, and $9 \%$ bandwidth, respectively. From this, one can observe that the stepped septum can better satisfy a broader bandwidth performance than that of sigmoid septums, and both sigmoid septums show excellent circularity within a narrow bandwidth. It can also be observed that the bandwidth for the $25 \%$ sigmoid septum is broader than the $5 \%$ sigmoid septum. As discussed above, the fitness functions of the two sigmoid septums are optimized within different bandwidths, which is reflected in the results of the simulations.


Figure 5.12: Simulated reflection coefficients for the coaxial port 1 as depicted in Figure
5.7.


Figure 5.13: Simulated isolation level between the two coaxial ports as depicted in Figure

In Figure 5.12, the simulated reflection coefficients for the three optimized horns are shown. The reflection coefficients are all around -20 dB for the three optimized horns. It is important to remember that these results were simulated with the coaxial probes included. Therefore the overall design including the horn, septum, and waveguide-tocoaxial probe demonstrate good reflection performance according to the widely used -10 dB impedance matching criterion. Next, the simulated isolation level between the two coaxial feed ports is shown in Figure 5.13. An isolation level below -20 dB is obtained between the two ports for all three designs. It is interesting to observe that the $5 \%$ sigmoid design achieved very good isolation $(<-35 \mathrm{~dB})$ over the bandwidth of interest. The PSO algorithm chose a design that by chance had a good isolation performance as well. The other designs also have fairly good isolation. Yet, if better isolation between the two ports is required, this isolation can also be included in the fitness function.


Figure 5.14: Radiation pattern for $25 \%$ sigmoid septum (RHCP: Co-pol, LHCP: X-pol).


Figure 5.15: Radiation pattern for $25 \%$ stepped septum (RHCP: Co-pol, LHCP: X-pol).


Figure 5.16: Radiation pattern for 5\% sigmoid septum (RHCP: Co-pol, LHCP: X-pol).

Figures 5.14-5.16 shows the simulated co-polarization (RHCP) and cross-polarization (LHCP) radiation patterns for the horn antennas loaded with the three optimized septums. The radiation patterns at 5.8 GHz are presented. The simulated radiation patterns provide a directivity of roughly $11 \mathrm{~dB}, 11.2 \mathrm{~dB}$, and 11.2 dB for the stepped septum, $5 \%$ sigmoid, and $25 \%$ sigmoid septum, respectively. The cross-polarization at boresight is around 30 dB less than the co-polarization for all three optimized horns, and the sidelobe levels are also very low. Another important point to remember is that these are the patterns when the upper port shown in Figure 5.7 and Figure 5.10 is excited. Similar patterns with opposite polarization are observed when the other port is excited.

### 5.4.2 Investigation of the Electric Field Intensity around the Stepped and Sigmoid Edges

The septum corners can induce a high electric field, resulting in air breakdown inside the horn antenna. Profiling the septum contour is critical to mitigate the air breakdown. In this section, the smooth and stepped profile of the septum are evaluated for handling HPM. The power handling of the horn antennas is greatly affected by the electric field intensity around the septum edges. The electric field around the edges is evaluated using HFSS, and the improvement provided by the sigmoid septum is presented versus that of the stepped septum. In Chapter 5.4.3, we test this approach in HFSS by simulating the canonical half-plane scattering problem. We then compare the HFSS simulation results with the analytical scattering solution around the edge. To our best knowledge, a septum with rounded edges has not been used to reduce the electric field around the septum. In Appendix II, the effect of the rounded edges on the electric field is discussed.


Figure 5.17: Electric field intensity evaluation on the septum plane using HFSS simulation.

Because the maximum electric field intensity inside a horn antenna is typically induced around the edge of the septum, our investigation was focused on the electric field near the edge. The simulated electric field intensity is evaluated on the septum plane that resides above the septum and extends along the waveguide section. We are interested in the electric field around the smooth and stepped edges when the same magnitude of incident waves illuminates the edges. From this evaluation, the relative improvement of the power handling can be accounted for in terms of the reduced maximum electric field intensity

Table 5.2 Maximum Total E-field and Each Component at the Location Where the Maximum Total E-field Occurs at 5.8 GHz with 1W Incident Power

|  | $\mathbf{E}_{\text {total }}(\mathrm{V} / \mathrm{m})$ | $\mathrm{E}_{\mathrm{x}}(\mathrm{V} / \mathrm{m})$ | $\mathrm{E}_{\mathrm{y}}(\mathrm{V} / \mathrm{m})$ | $\mathrm{E}_{\mathrm{z}}(\mathrm{V} / \mathrm{m})$ |
| :---: | :---: | :---: | :---: | :---: |
| $25 \%$ STEPPED | 15331 | 8850 | 741 | 12496 |
| $25 \%$ SIGMOID | 8590 | 4533 | 1256 | 7187 |
| $5 \%$ SIGMOID | 6527 | 3516 | 2315 | 4987 |

around the edges. Since the maximum electric field is usually seen in a small region, setting a fine mesh to the HFSS model is important for properly evaluating the electric field. In this work, a fine meshing is applied to the surfaces of interest, which is the septum plane residing above the septum polarizer as shown in Figure 5.17. This setup is helpful to reduce the computational size of the HFSS model even though fine meshing in the entire volume is desirable. Figure 18 shows the setup of the mesh along the septum plane, illustrating the triangular shape of a single mesh at 5.8 GHz . Each side of the triangles in the septum plane is limited to $\lambda / 50$ in this investigation for the sigmoid and stepped septum polarizers. This mesh size is also applied to the half-plane scattering provided in Chapter 5.4.3. A 1W incident wave is excited in the circular waveguide by using a waveport in HFSS, and the induced electric field along the septum plane is observed.

The simulated electric field intensity along the septum plane is shown in Figures 5.185.20 for all three septum designs. The magnitude of the electric field intensity along a
waveguide section is also magnified where the maximum electric field is observed. The maximum of the total electric field and each electric field components $E_{x}, E_{y}$, and $E_{z}$ for the total field are listed in Table 5.2. It is shown that a significant reduction of the electric field can be achieved using the sigmoid septum. For the sigmoid septums, the maximum electric field components including the maximum total electric field is reduced by $30 \%$ $50 \%$, in comparison to the stepped septum. In addition, we can compare the maximum of the electric field components between the $5 \%$ and the $25 \%$ sigmoid septums. Total electric field components of the $5 \%$ sigmoid septum are $24 \%$ less than those of the $25 \%$ sigmoid septum because the $5 \%$ sigmoid septum has a smoother septum profile compared to the $25 \%$ sigmoid septum due to the higher value for the C variables, as discussed in Chapter 5.4.1. In summary, the maximum of the total electric fields around the edges can be listed in an order of $25 \%$ stepped septum, $25 \%$ sigmoid septum and $5 \%$ sigmoid septum. The $25 \%$ sigmoid and $5 \%$ sigmoid septums can reduce the maximum of the total electric field by $43 \%$ and $57 \%$, compared to the maximum fields of the stepped septum, respectively. Following the relationship of $\mathrm{P} \propto|\mathrm{E}|^{2}$, the power handling can be enhanced by roughly 3 times for both sigmoid septums.

Visualization of the electric field also provides a physical understanding for the electric field near the contour of the septum. Note that Ex and Ey are tangential components to the septum plane, while Ez is the normal component to the septum plane. Based on the PEC boundary condition and the analysis of half-plane scattering in 5.4.3, the electric field distribution around the edge can be described as follows:

1) Only the Ez component exists above the septum.
2) The Ex and Ey components are observed outside the septum. Only the components normal to the edge can exist while no components can exist parallel to the edge.
3) The maximum of the total electric field is induced near the contour of the septum.

The solution to the half-plane scattering problem [26] also indicates that a hard or soft polarized incident wave generates a maximum of the total electric field near the edge. To evaluate the numerical approach, the simple half-plane scattering is compared with the analytical solution as shown in the Appendix. Good agreement is observed between the numerical and analytical solutions, which verifies this technique to determine the maximum electric field.

These physical principles and observations from the canonical half-plane scattering problem can be observed in the results shown in Figures 5.18-5.21. Indeed the existence of PEC in a given region forces the electric field components tangential to the metal surface to be zero due to the boundary conditions. We also observe that the maximum electric field occurs in the region near the edge, as shown on the inset figures. These observations provide both reassurance into the computational technique and a better physical understanding of the septum. We have also claimed that the sigmoid is able to handle wide bandwidths in terms of its RF performance in the previous section. Therefore, one must also demonstrate good power handling capabilities over the entire bandwidth of interest in order to fully claim that this design can obtain a wideband performance with good power handling.


Figure 5.18: Total electric field intensity on the septum plane and magnified visualization of each field component at the location where the maximum total electric field occurs at
5.8 GHz with 1 W incident power.


Figure 5.19: $\mathrm{E}_{\mathrm{x}}$ electric field intensity on the septum plane and magnified visualization of each field component at the location where the maximum total electric field occurs at 5.8

GHz with 1 W incident power.


Figure 5.20: $\mathrm{E}_{\mathrm{y}}$ field intensity on the septum plane and magnified visualization of each field component at the location where the maximum total electric field occurs at 5.8 GHz with 1 W incident power.


Figure 5.21: $\mathrm{E}_{\mathrm{z}}$ field intensity on the septum plane and magnified visualization of each
field component at the location where the maximum total electric field occurs at 5.8 GHz with 1 W incident power.

We used the same procedure outlined in the previous paragraphs and computed the maximum electric fields for several frequencies within the bandwidth. The results of these simulations are shown Figure 5.22, and clearly one can observe that both sigmoid septums perform better than the stepped septum in terms of power handling. Note that the maximum field location might be at another location for different frequencies.

It is also interesting to observe that the $5 \%$ sigmoid has a lower maximum electric field for all frequencies in comparison to the $25 \%$ sigmoid. This was predicted in Section 3.3 when we showed that the average C variables for the $5 \%$ design were higher than the $25 \%$ design, leading to a smoother curve. Overall, the results meet our expectations in terms of power handling capabilities.


Figure 5.22: Electric field intensity evaluation on the septum plane using HFSS simulation when 1 W is applied to the coaxial probe.

The effectiveness of the sigmoid septum for use in HPM systems is demonstrated with the reduced electric field intensity near the edge. While a small loss of the AR/S11 bandwidth is experienced for the sigmoid designs, they are still attractive in terms of handling HPM. In essence, there is a tradeoff between bandwidth and power handling capability that a designer must make when working with these septum designs.

## 5. 4.3 An HFSS Evaluation of the Half-plane Scattering Problem



Figure 5.23: Geometrical configuration of the half-plane for both soft and hard polarizations.

In order to test the numerical approach in simulating the electric field intensity around the edge of the septum, a half-plane scattering for the hard and soft polarization is simulated with HFSS and is compared with an exact analytical solution. The current distribution and electric field distribution along the half-plane are investigated for both polarizations.

Figure 5.23 describes the scattering scenario of interest. The edge is located along the z axis, and the edge is illuminated by either a hard or soft polarization incident plane-wave at the angle $\varphi_{0}=90^{\circ}$. The distance $\rho$ is measured from the edge along the z axis, and the angle $\varphi$ is defined from the half-plane to the observation point. For this investigation, we
will examine the current density and the electric field distribution along the x -axis since the maximum electric field occurs along this plane. With this, the distance $\rho$ and angle $\varphi$ are also measured along the x -axis.

Next, an analytical model is needed for the comparison. The series solution presented in [26] is used to obtain the current density along the x -axis for both polarized incident waves. For the scattering scenario discussed above, the expression of the current density due to the hard polarization incident wave can be reduced to

$$
\begin{gather*}
J_{\rho, \text { hard }}^{\text {Too }}=H_{0} \sum_{n=0}^{\infty} \varepsilon_{n} j^{n / 2} J_{n / 2}(k \rho) \cos \left(\frac{n \pi}{4}\right)  \tag{5.6}\\
\text { (Top of half-plane) } \\
J_{\rho, \text { hard }}^{\text {Bottoo }}=-H_{0} \sum_{n=0}^{\infty} \varepsilon_{n} j^{n / 2} J_{n / 2}(k \rho) \cos \left(\frac{n \pi}{4}\right) \cos (n \pi)
\end{gather*}
$$

(Bottom of half-plane)

$$
\begin{equation*}
J_{\rho, \text { hard }}^{\text {Total }}=J_{\rho, \text { hard }}^{\text {Top }}+J_{\rho, \text { hard }}^{\text {Botom }} \tag{5.8}
\end{equation*}
$$

(Total current density)
where $\varepsilon_{n}\left\{\begin{array}{ll}1 & n=0 \\ 2 & n \neq 0\end{array}(n=0,1,2 \ldots)\right.$ and $J_{n / 2}(k \rho)$ is the Bessel function of the first kind of order $\mathrm{n} / 2$. Note that the angle $\varphi$ was substituted by $\varphi=0^{\circ}$ and $\varphi=360^{\circ}$ to obtain the current on the top and bottom of PEC half-plane, respectively. The total electric field due
to the hard polarized incident waves can be written by the following equations.

$$
\begin{gather*}
E_{\rho, \text { hard }}^{\text {Total }}=-\frac{H_{0}}{j \omega \varepsilon \rho} \sum_{n=0}^{\infty} \varepsilon_{n} j^{n / 2} \frac{n}{2} J_{n / 2}(k \rho) \cos \left(\frac{\pi}{4} n\right) \sin \left(\frac{n}{2} \varphi\right)  \tag{5.9}\\
E_{\varphi, \text { hard }}^{\text {Total }}=\frac{k H_{0}}{j \omega \varepsilon} \sum_{n=0}^{\infty} j^{n / 2} J_{n / 2}^{\prime}(k \rho) \cos \left(\frac{\pi}{4} n\right) \cos \left(\frac{n}{2} \varphi\right)  \tag{5.10}\\
\vec{E}_{\text {hard }}^{\text {Total }}=\hat{\rho} E_{\rho, \text { hard }}^{\text {Total }}+\hat{\phi} E_{\varphi, \text { hard }}^{\text {Total }} \text { (Total electric field) } \tag{5.11}
\end{gather*}
$$

where $J_{n}^{\prime}(k \rho)=\frac{\partial}{\partial(k \rho)}\left[J_{n}(k \rho)\right]$ in equation (5.10). It should be noted that the equations for the hard polarization case are given for an incident plane wave which can be written as $H_{z}^{i}=H_{0} \exp (j k y)$ [27].

The current density due to the soft polarization incident plane wave can be reduced to

$$
\begin{equation*}
J_{z, \text { soft }}^{\text {Top }}=\frac{E_{0}}{j \omega \mu \rho} \sum_{n=1}^{\infty} j^{n / 2} J_{n / 2}(k \rho)\left(\frac{n}{2}\right) \sin \left(\frac{n \pi}{4}\right) \tag{5.12}
\end{equation*}
$$

(Top of half-plane)

$$
\begin{equation*}
J_{z, \text { soft }}^{\text {Botom }}=-\frac{E_{0}}{j \omega \mu \rho} \sum_{n=1}^{\infty} j^{n / 2} J_{n / 2}(k \rho)\left(\frac{n}{2}\right) \sin \left(\frac{n \pi}{4}\right) \cos (n \pi) \tag{5.13}
\end{equation*}
$$

(Bottom of half-plane)

$$
\begin{equation*}
J_{z, \text { soft }}^{\text {Total }}=J_{z, \text { soft }}^{\text {Top }}+J_{z, \text { soft }}^{\text {Botom }}(\text { Total current density }) \tag{5.14}
\end{equation*}
$$


High electric intensity
Hard-polarization incident wave


## Soft-polarization incident wave

Figure 5.24: Simulated electric field on the observation plane for both hard and soft polarization.

The series solution of the total electric field intensity for the soft polarization incident plane wave can be written as

$$
\begin{equation*}
E_{z, \text { soft }}^{\text {Toal }}=E_{0} \sum_{n=1}^{\infty} j^{n / 2} J_{n / 2}(k \rho) \sin \left(\frac{n \pi}{4}\right) \sin \left(\frac{n \varphi}{2}\right) \tag{5.15}
\end{equation*}
$$

These equations are given for an incident plane wave of the form $E_{z}^{i}=0.5 E_{0} \exp (j k y)$, where $E_{0}$ is a constant that is arbitrary.

The simple half-plane model is setup in HFSS in the same manner as the septum for the purpose of evaluation. Figure 5.24 depicts the setup of the half-plane model in HFSS. Although the infinite half-plane is ideal for this evaluation, a finite width strip is implemented into HFSS due to the limited numerical resources. The width of the halfplane is $20 \lambda$ at 5.8 GHz . Note that the large width allows the use of the finite strip due to minimal interactions between the two edges. The current density and electric field intensity are obtained along the observation plane across the edge. For a fair verification, a similar mesh is implemented in the half-plane model, where each mesh element edge is set to $<\lambda / 50$.

The comparison between the series solution and simulation is made for the current density and electric field intensity. Figure 5.25 and Figure 5.26 shows the comparison of the current density along the observation plane. Note that the simulated current densities are normalized to the magnitude of the $H_{0}$ and the electric field intensity is normalized to the magnitude of the $E_{0}$. The simulated current densities for both polarizations are in good agreement with the series solution.


Figure 5.25: Simulation and series solution of the current density for hard-polarized incident wave.


Figure 5.26: Simulation and series solution of the current density for soft-polarized incident wave.

Figure 5.27 and Figure 5.28 depicts the comparison of the electric field intensities on the observation plane, which shows a good agreement between analytical solutions and simulated results. In particular, a close similarity can be seen with respect to the magnitude of the maximum field intensity and position where the maximum occurs. The electric field intensity for the hard polarized incident wave is shown in Figure 5.26, which demonstrates a high magnitude near the half plane edges. For the soft polarization scenario shown in Figure 5.28, the series solution agrees well with the simulated electric field intensity.


Figure 5.27: Simulation and series solution of the electric field intensity for hardpolarized incident wave.


Figure 5.28: Simulation and series solution of the electric field intensity for soft-polarized incident wave.

## 5. 4.4 Comparison to a Septum with a Round Chamfered Septum

There are still several questions one might ask when approaching this engineering design problem. One might ask how much improvement could be expected when comparing the stepped septum with sharp edges (traditional septum design) to a stepped septum with round chamfered corners. Figure 5.29 shows the configuration of a possible round chamfered septum and its maximum total E-field comparison between the stepped septum and the round chamfered septum. The distribution of each component is shown in Figure 5.30. Table 5.3 demonstrates a reduction of the round corner septum in comparison to the sharp corner septum, and it is shown that the maximum total E-field of the round chamfered septum can be reduced to the one of the sigmoid septum that we investigated in Section 5. Therefore, smoothing the edges in this transverse direction also reduces the electric fields.

However, there are two major problems with smoothing performed in this direction. The first is that it makes the fabrication much more difficult to maintain good tolerances on the septum contour. The second issue is that the benefit of rounding the corner is very similar to the smoothing performed by the sigmoid function. Smoothing in the transverse direction does indeed improve the power handling, but the cost far outweighs the benefits.


Max total E-field comparison between 25\% stepped and rounded stepped septum

Figure 5.29: Configuration of the round chamfered septum and total maximum E-field of the stepped septum and round chamfered septum at 5.8 GHz with 1 W incident power.


Figure 5.30: Electric field components on the plane of the round chamfered septum.

Table 5.3 Maximum Total E-field and Each Component at the Location Where the Maximum Total E-field Occurs at 5.8 GHz and 1W Incident Power

|  | $\mathbf{E}_{\text {total }}(\mathbf{V} / \mathbf{m})$ | $\mathrm{E}_{\mathrm{x}}(\mathrm{V} / \mathrm{m})$ | $\mathrm{E}_{\mathrm{y}}(\mathrm{V} / \mathrm{m})$ | $\mathrm{E}_{\mathrm{z}}(\mathrm{V} / \mathrm{m})$ |
| :---: | :---: | :---: | :---: | :---: |
| $25 \%$ <br> STEPPED | 15331 | 8850 | 741 | 12496 |
| $25 \%$ <br> ROUND <br> CHAMFER <br> ED | 7812 | 3516 | 574 | 6953 |

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## Chapter 6

## Conclusions

This dissertation has presented some recent progress for antennas in wireless communications and near-field linkbudget analysis. In particular, a new formulation for near-field analysis has been given to evaluate the link between two antennas when in close proximity (in the so-called near field region). Several new antenna designs with various functionalities were also introduced for many different applications including wireless communications. The first implements frequency reconfigurability using RFMEMS switches. The second adds electronic beam tilting using electromagnetic bandgap (EBG) structures. The last uses Particle Swarm Optimization to provide a smooth contour for a septum design which creates circular polarization from a linear polarization. Accomplishments in this dissertation can be summarized in detail as shown below:

- Comprehensive evaluation and simplification of the near-field coupling problem: Two radiating near-field coupling formulas: 1) generalized Friis formula and 2) integral coupling formula have been investigated, and their validity has been shown by evaluating the link budget for several antennas using full-wave simulations and measurements. Firstly, a generalized Friis formula is investigated for possible simplifications which lend to a more practical use. A simplified correction term to
the Friis formula is presented to accurately estimate the link budget between two antennas within near-field region. The proposed correction term provides a significant improved accuracy in the Fresnel region and requires only simple calculations. For non-bore-sight scenario, a cosine approximation of the radiation pattern in one plane is proposed. Next, the integral coupling formula is comprehensively investigated. The formula is effective in the entire range of the near field region, and more advances are made to calculate the coupling in a variety of antenna configurations. The power transmissions of a variety of antennas are evaluated. An improvement of $2-10 \mathrm{~dB}$ is observed in the prediction made by the formula compared to the standard Friis formula. Indoor measurements also validate the effectiveness of the proposed method.
- Design implementation of MEMS frequency reconfigurable patch antenna: The design of the patch antenna with switchable slot (PASS) is implemented with a commercially available RF MEMS switch and a bias network. Unique DC bias lines are integrated into the slot structure to actuate the switch placed in the slot. The inclusion of the dc bias lines necessitates additional circuit components for the proper operation of the bias network. The filter designs include an inter-digital capacitor and spiral inductor, which are integrated into the bias network. More importantly, the influence of the MEMS switch and bias network on the antenna performance was investigated. It has been demonstrated that adding a MEMS switch and wire-bonding has minor effects on the radiation mechanism of the patch. The proposed design has been fabricated, and it has been measured for both off and on states. The antenna operates at two different frequency bands, which can be chosen
by the MEMS switch state. It was observed that the measured pattern is similar to the typical pattern of the patch antenna. The proposed design can be applicable to many slot loaded structures and can also be expanded to phased arrays.
- Beam-tilting element in the array and EBG base-station: Traditional array antennas are vulnerable to gain loss when the beam is steered toward a tilted direction. The beam tilting element incorporated into an array antenna can compensate the gain loss of the array antenna. To verify this concept, the beamtilting EBG-dipole is designed as a single element, which can be extended as an array antenna for base-station applications. The beam-tilting EBG-dipole is realized with a dipole antenna mounted on the tunable EBG ground. In addition, another EBG-dipole antenna suitable for a mechanically adjusted base-station is also designed. A higher gain, compared to the base-station antennas is achieved across the bandwidth of interest, and a reasonable impedance matching of the proposed antenna has been obtained. A link-budget evaluation has been performed to observe the enhancements between a mobile and its corresponding base-station.. Significant enhancements in base-station coverage are obtained by utilizing the proposed array antenna. Therefore, a high performance and robust array antenna has been realized with the proposed antenna.
- Design and Optimization of a Smooth septum for HPM horn antenna design: A smooth profile septum placed inside a circular horn was presented for possible use in high power microwave systems. The fundamental operation of the septum was revisited using graphical evaluation. The Particle Swarm Optimization (PSO)
technique was implemented to optimize the smooth septum geometries profiled with the sigmoid function as well as the stepped septum. The optimized geometry set for the sigmoid and stepped septums was provided by PSO, and good RF performance was observed in each of these final designs. In addition, the electric field around the septum is investigated in order to examine the power handling of the optimized sigmoid and stepped septums. It was verified that roughly 3 times improvement of power handling can be achieved from the sigmoid septum. With this sigmoid septum, the usable power range for many applications can be greatly extended, accompanied with a good circularity over a broad bandwidth.

