Ka-band 2D Luneburg Lens Design with Glide-symmetric Metasurface

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Abstract

A Luneburg lens is a beam former that has two focal points where one is at the surface and the other lies at infinity. It is a cheap passive steerable antenna at high frequencies. In this thesis, a 2D flat-profile Luneburg lens with all-metal structure is designed for Ka band. Commercial software CST Microwave Studio Suite and Ansys Electronic Desktop (HFSS) are used for simulations.

The lens is composed of two glide-symmetric metasurface layers with a small gap in between. The high order symmetry, glide symmetry, could provide ultra wide band property for the lens. Each layer contains many unit cells. Different unit cells are tested in this thesis to find the best solution taking into account both electromagnetic properties and the easiness of manufacturing. A flare is designed to achieve better matching between the air gap of the lens and free space. A self-designed waveguide feeding is also used, including a transition from coaxial cable to TE_{10} mode of rectangular waveguide at the focus of the lens.

The prototype will be built in the future and measurements will be done to compare with simulation results in this thesis.
Abstract


En prototyp kommer att byggas i ett senare skede och mätningar göras för att jämföra med simuleringsresultaten i detta examensarbete.
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Nomenclature

Abbreviations

EBG Electromagnetic Bandgap
EDGE Enhanced Data Rates for GSM Evolution
Gbps Gigabits Per Second
GPRS General Packet Radio Service
GSM Global System for Mobile Communications
HSPA High-Speed Packet Access
IMT International Mobile Telecommunications
IP Internet Protocol
ITU Internet Telecommunication Union
kbps Kilobits Per Second
LTE-A Long Term Evolution Advanced
LTE Long Term Evolution
Mbps Megabits Per Second
MIMO Multiple Input Multiple Output
OFDM Orthogonal Frequency-division Multiplexing
PEC Perfect Electric Conductor
PMC Perfect Magnetic Conductor
TEM Transverse Electric and Magnetic Field
UTMS  Universal Mobile Telecommunication System
WiMAX Worldwide Interoperability for Microwave Access

Notations

- $\epsilon_r$  Relative permittivity
- $\lambda$  Wavelength
- $c$  Speed of light in vacuum
- $n$  Refractive index
- $n_{\text{eff}}$  Effective refractive index
Chapter 1

Introduction

1.1 Background Information

The 1G (first generation of wireless mobile communication technology) was introduced in the 1980s with analog cellular networks. This was first commercially used in Japan by NTT (Nippon Telegraph and Telephone) for public voice service in 1979. Later in 1990s, the 2G (second generation mobile phone technology) based on digital transmission emerged. It primarily used the GSM (Global System for Mobile Communications) standard deployed first in Finland by Radiolinja in 1991. The 2G systems were significantly more efficient in the phone-to-network signaling with digital coding. They provided better voice clarity and encryption for phone calls, and introduced SMS (Short Message Service) text messaging as well as data surfing for mobile phones. Later, 2.5G and 2.75G wireless cellular networks were developed along the way to 3G. Two main examples are GPRS (General Packet Radio Service) and EDGE (Enhanced Data Rates for GSM Evolution). GPRS could provide from 56 kbps up to 115 kbps data rate, while the speed for EDGE was up to 384 kbps.

With the explosive success of 2G mobile phones, the demand for data services grew hugely. Then 3G high speed network technology using packet-switching came into being since 1998. One example is UMTS (Universal Mobile Telecommunication System) originated in Europe. It allows from 384 kbps to 2 Mbps data speed for different mobility and coverage. It is also the base of the HSPA (High-Speed Packet Access) protocol, which is sometimes called 3.5G, and has a maximum speed of 14 Mbps.

The 4G technology is an all IP (Internet Protocol)-based network system. It treated voice calls just like a kind of streaming audio media, and eliminated circuit switching. 4G is expected to integrate all existing and future wireless networks and provide
broadband as well as smooth global roaming. The first commercially deployed standard is LTE (Long Term Evolution) in Stockholm and Oslo in December 2009. The later technologies LTE-A (Long Term Evolution Advanced) and WiMAX (Worldwide Interoperability for Microwave Access) can provide up to 100 Mbps or 1 Gbps data transfer speed depending on mobility, while the latency is around 70 milliseconds.

And now we are on our migration to 5G. The final 5G standard IMT(International Mobile Telecommunications)-2020 will be released by ITU (International Telecommunication Union) standards body. The goal for 5G is to achieve 20 Gbps speed and 1 millisecond latency with about 1000× higher capacity than LTE network. Users should be able to download high definition films in less than a second, while the task time needed for 4G LTE is 10 minutes. With such powerful network, smart house, autonomous vehicles, and the Internet of Things etc. could become true. However, such massive capacity could only be achieved with very high operating frequencies, such as millimeter waves. But high frequency electromagnetic waves will imply close propagation distance due to increased path loss and low wall-penetration capability. Thus, small size 5G antennas are required to be relatively close to each other. One solution is to build Small Cells, or even home routers, instead of huge towers. Also, in order to promote the antenna performances, 5G will probably take advantage of OFDM (Orthogonal Frequency-division Multiplexing) encoding as well as Massive MIMO (Multiple Input Multiple Output) technology.

Massive MIMO is to operate a large number of 5G service antennas coherently in an array, thus increase the network capacity by a large factor. But the more users there are, the more they will interfere with each other. Hence, beamforming and cancellation (“nulling”) are necessary technologies for 5G base stations. They help focusing signals in a concentrated beam that is very directive, and reducing signal blockage and weakening in unnecessary directions, thus decrease interference among different users.

Some interesting 5G frequency channels are 24 GHz, 28 GHz, and 39 GHz. The goal of this thesis is to design a high-efficiency multibeam Luneburg lens antenna that could cover a large angular range and provide wide bandwidth at around 28 GHz.

Apart from 5G, Luneburg lens can also find application in radio link systems, especially in mesh networks. A radio link is a wireless point-to-point connection between 2 nodes in a network, where every node includes a transceiver and a very directive
antenna. Two nodes will be mounted pointing towards each other with no obstacles in between, and the directive antennas make sure a high data rate.

1.2 Luneburg Lens

A Luneburg lens is a spherically symmetric lens with gradient refractive index. In 1944, R.K Luneburg developed the basic theory of this lens [1]. It is a beam former that has two focal points, one at the outer surface, and the other at infinity in the opposite direction. That is to say, if we feed with a point source at the lens surface, we get a plane wave on the other side, as shown in Figure 1.1, and vice versa. A perfect Luneburg lens is isotropic and could support full angular coverage by moving the feeding point. At high frequencies, this kind of lens can be a feasible steerable antenna since arrays and phase shifters are costly and complex [2].

A Luneburg lens follows a relation between refractive index and radial position given in Equation 1.1. In this equation, $\rho$ stands for the current radius while $R$ represents the out-most radius of the Luneburg lens. It can be seen from Figure 1.2 that the refractive index $n$ falls from $\sqrt{2}$ to 1 from center to surface, while the relative permittivity $\epsilon_r$ varies from 2 to 1. If a Luneburg lens is placed in air, there will be no reflections at the lens-air interface since the refractive index at lens surface is the same as that of air. This solves the problem of reflection with conventional lenses.

$$n = \sqrt{\epsilon_r} = \sqrt{2 - \left(\frac{\rho}{R}\right)^2} \quad (1.1)$$
1.3 Antenna Design Specifications

Traditionally, Luneburg lenses are 3D structures made with dielectrics [3] [4]. They are expensive due to complex manufacturing processes, and also quite bulky as well as lossy to use in some practical applications. In this thesis, I explored a cost-effective planar Luneburg lens configuration with glide-symmetric metasurface, to achieve a low-profile, high gain, low loss and broadband antenna with wide-angle scanning capability. It is meant for 5G radio-link applications, but also has potentials to be investigated more deeply and used in other Ericsson wireless communication applications. The whole design took into account manufacturing, assembly and real-life cost. A prototype will be manufactured afterwards. The detailed specifications are listed in Table 1.1.

<table>
<thead>
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<th>Table 1.1: Project Specifications</th>
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<tbody>
<tr>
<td>Center frequency</td>
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<td>Bandwidth</td>
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<td>Beam width</td>
</tr>
<tr>
<td>Scan capability</td>
</tr>
<tr>
<td>Others</td>
</tr>
</tbody>
</table>
1.4 The State of Art

Since Rudolf Luneburg proposed the simple solution to generate two special foci with the gradient-index lens, different methods have been investigated to manufacture and implement it in microwave antenna applications. In 1958, G. Peerler and H. Coleman showed that 10 discrete layers of dielectric were sufficient to provide Luneburg lens behavior instead of continuously varying refractive index materials [3]. Their lens had 18 inches diameter and was aimed to be used at X band. They adopted equal $\epsilon_r$ increment method for the 10 layers instead of equal $n$ increment, as relative permittivity is usually more directly measured than refractive index.

S. Baev et al. further explored the focusing effect of multi-layer dielectric Luneburg lens with frequencies. They simulated a 10 layer lens with 200 mm diameter with wave frequency varying from 2.45 GHz to 10 GHz [4]. They showed that splitting the 3D lens model into a $\frac{1}{4}$ sphere is enough to get correct radiation pattern in simulation due to E-field symmetry and H-field symmetry.

In 1952, G. Peeler and D. Archer also built a 2D Luneburg lens operating in TE$_{10}$ mode with almost-parallel plate waveguide filled by polystyrene [5]. They tuned the refractive index by changing the thickness of polystyrene. In [6–8], the method of controlling permittivity changed to drilling small holes in dielectric disks instead of varying the height of dielectric. Adding air holes can decrease the equivalent dielectric constant by having less material per unit volume, and the density of holes determines the resulting refractive index.

However, all these lenses have quite complex manufacturing procedures, like binding dielectric layers, contouring dielectrics, or drilling numerous holes.

To achieve easier fabrication, C. Pfeiffer and A. Grbic designed a 2D metasurface Luneburg lens with printed circuit board technique in 2010 [2]. They etched crossed microstrip lines over a copper-clad substrate and controlled refractive index by meandering the transmission lines and varying line widths on each unit cell. Their lens operated at 13 GHz in TEM mode, with tapered transmission line feed and a flare. The scanning angle was from $-45^\circ$ to $+45^\circ$ with -3 dB cross over level. Cheng et al. also made use of etching technique to fabricate a Luneburg lens with I-shaped unit
cells at around 10 GHz [9].

Also in 2010, M. Casaletti et al. presented another way to achieve 2D Luneburg lens refractive index profile using Fakir’s bed of nails substrate [10] inside a parallel plate waveguide. They modified the pin height to achieve the needed local refractive index at 10 GHz [11].

In 2012, M. Bosiljevac et al. applied a circular patch array with varying patch sizes on a dielectric substrate, and achieved Luneburg lens behavior inside a parallel plate waveguide [12]. The center frequency of their design is 13 GHz and the dielectric had a permittivity of 10.2. They also suggested two different flare structures to match the impedance of thin waveguide and free space. However, since most of the energy was confined inside the substrate, the effect of flares were actually limited. In [13], a similar configuration was proposed.

Although huge progress has been made in fabricating methods, all these Luneburg lenses contain dielectric. At high frequencies, electromagnetic waves propagating along the surface of dielectric substrate will extend many wavelengths into dielectric, causing relatively high dielectric loss. Moreover, some of them do not support TEM wave, which leads to intrinsically limited bandwidth [14].

As a contrast, metals are good conductors with large and imaginary dielectric constant at microwave frequencies. Electromagnetic waves are almost all screened out due to high conductivity of the metal, stopping fields to propagate inward. Recently, a full metal metasurface with 2 layers of glide-symmetric configurations was proposed by O. Quevedo-Teruel et al. [15]. It has higher effective refractive index and ultra-wideband, resulting in a flat Luneburg lens operating from 4 to 18 GHz. In order to implement this technology at high frequency (60 GHz), a different unit cell for Luneburg lens was proposed by A. Torki et al. in [16]. Some other applications of glide symmetry technology can be found in [17–19].

1.5 Thesis Outline

In Chapter 2, the basic theory of a glide-symmetric metasurface, and the designing principle of unit cells are explained. In Chapter 3, four different unit cells that have been investigated are presented, as well as the final choice for the metasurface lens.
Then, Chapter 4 explains the optimum lens layout considering both dispersion and easiness of manufacturing. Chapter 5 deals with the design of flare to match the impedance of the lens gap and free space, and the design of a waveguide feeding to provide best matching and integral configuration. In Chapter 6, the overall simulation results obtained by using CST Microwave Studio Suite, and the scanning properties of the lens are shown. Chapter 7 evaluates some critical tolerances on the manufacturing. Chapter 8 draws the conclusions and gives future perspectives.
Chapter 2

Theory

2.1 Metasurface

An electromagnetic metasurface refers to a surface formed by artificial materials with boundary conditions not found in nature. They can be applied to produce unusual reflection/refraction properties of incident waves and can also guide surface waves in 2D configurations. Typical metasurfaces include PEC (Perfect Electric Conductor), AMC (Artificial Magnetic Conductor, emulating the effect of Perfect Magnetic Conductor that does not exist in nature), EBG (Electromagnetic Bandgap) surfaces, Soft and Hard surfaces etc. Some typical configurations are periodic arrays of patches or holey metallic structures over a dielectric substrate [20], and metallic pins [21] [10]. In this thesis, PEC and PMC’s particular characteristics were taken advantage of in designing the specific metasurface with Luneburg lens performance.

Perfect Electric Conductor:
Perfect electric conductor is an idealized material of metal. It has infinite electrical conductivity, i.e. no resistivity, hence no heat will be generated inside. The PEC canonical surface is commonly used in antenna analysis for metal conductors when electrical resistance is negligible compared to other effects. Along PEC surfaces, only vertically polarized electromagnetic waves can propagate. Parallelly polarized waves will be stopped.

Perfect Magnetic Conductor:
The PMC canonical surface is usually used in theoretical works and as symmetry plane for design purpose. It can be considered as a reciprocal of PEC. Electromagnetic waves with magnetic field parallel to the surface do not propagate, only those with perpendicular magnetic field can pass. In other words, only parallelly polarized
waves can propagate along PMC surfaces.

Metasurfaces designed for antennas usually have sub-wavelength thickness [22] and contain sub-wavelength periodic repetition of unit cells [15]. Using the periodic structure of metasurface, it is possible to spatially vary its electromagnetic responses by modifying the dimensions of unit cells. Thus, a grade index feature on a flat surface is achievable.

### 2.2 Glide Symmetry

Glide symmetry is a higher order symmetry compared to simple translation or reflection. It applies to periodic structures when they stay invariant under glide operation $G$. That is, a translation of half period $p$ along the glide plane, and a reflection over the plane [23]. In the Cartesian coordinate system of our model, the glide operation $G$ can be expressed as shown in Equation 2.1.

$$G = \begin{cases} 
  x \to x + p/2 \\
  y \to y + p/2 \\
  z \to -z 
\end{cases} \quad (2.1)$$

In Figure 2.1, the relation $G$ is drawn in a 2D sketch with Ericsson logo. There is a mirroring over $x$-$y$ plane, and a half period shift between two layers in both $x$ and $y$ direction. Taking advantage of this higher order symmetry, frequency dispersion can be greatly reduced and we can achieve ultrawideband property for metasurface [15, 24, 25]. The analyses of glide-symmetric corrugated structures have been done with equivalent-circuit method [26] and mode matching method [27, 28].

![Figure 2.1: Illustration of glide symmetry.](image)
2.3 Unit Cell Properties

One of the most interesting properties for metamaterial is the dispersion diagram, on which we can see the relation between phase and frequency, as well as the bandgap behavior for EBG structures. In CST Microwave Studio Suite, we use Eigenmode Solver to simulate a unit cell in an infinite periodic structure. The boundary conditions in the metasurface plane are "periodic", where we can set the phase shift in both $x$- and $y$- directions between adjacent unit cells. The boundary conditions on top and bottom ($z$- direction) are PEC ($E_t = 0$).

Figure 2.2 illustrates a typical unit cell of a periodic metasurface without glide symmetry. All boundary conditions are applied as demonstrated above. The $x$- and $y$- direction phase delays are represented by parameters "phaseX" and "phaseY" in CST. And we can cover the whole irreducible Brillouin zone by varying them.

![Figure 2.2: A typical unit cell in periodic metasurface.](image)

2.3.1 Brillouin Zone

In transmission lines, the propagation constant $\gamma$ is composed of two terms, attenuation constant $\alpha$ and phase constant $\beta$.

$$\gamma = \alpha + j\beta$$  \hspace{1cm} (2.2)

Originally dispersion diagram is the function between $\beta$ and frequency $\omega$, but in lossless case, $\alpha = 0$ and phase constant $\beta$ equals the wavenumber $k$ for a plane wave. In free space, the relation is linear:

$$\beta(\omega) = k_0 = \omega \sqrt{\mu_0 \varepsilon_0}$$  \hspace{1cm} (2.3)
However, for surface waves propagating in EBG structures, it is usually hard to give an explicit expression for $k$. Eigenvalue equations are usually solved or full wave simulations are performed in order to get the wave number. As is known, there might exist more than one solution for an eigenvalue equation, which means, the propagation constant might not be unique at one frequency. These different solutions are called modes. Each mode has its own field distribution, phase velocity and group velocity. The relation between $\beta$ and $\omega$ is often plotted out in curves called dispersion diagram.

In a periodic structure, the field distribution is also periodic with a phase delay between unit cells. This phase delay is determined by phase constant $\beta$ and periodicity $p$. For a surface wave mode propagating in $x$-direction in an infinite plane, its field can be written in a series of space harmonic waves:

$$
\vec{E}(x, y, z) = \sum_{n=-\infty}^{\infty} \vec{E}_n(y, z) e^{-j\beta_n x} e^{j\omega t}
$$

$$
\beta_n(\omega) = \beta(\omega) + n
$$

It’s clear that the periodicity of $\beta$ is $2\pi/p$. Therefore, the dispersion diagram only needs to be plotted in one period known as Brillouin zone. For our 2D surface, this is a square region where $0 \leq \beta_{xn} \leq 2\pi/p_x$ and $0 \leq \beta_{yn} \leq 2\pi/p_y$. And by symmetry, it can be further reduced to the irreducible Brillouin zone, whose edge is marked by the orange triangle in Figure 2.3, since the zone depends on the lattice. With $\Gamma X$, $XM$ and $M\Gamma$ track, we could simulate a surface wave propagating in $0^\circ$, $90^\circ$ and $45^\circ$, which set the limits for waves propagating in other directions.
In our case, the unit cell is an isotropic square in $x$-$y$ plane, hence the performance in $x$- and $y$- direction should be identical. As for $45^\circ$ wave, it is different with waves along the side in this example, but in a glide-symmetric unit cell, there is only tiny dispersion between waves in $0^\circ$ and $45^\circ$ for mode 1 [15]. Hence, we will only consider the dispersion diagram in $x$- direction (from $\Gamma$ to $X$) for later unit cell design with glide symmetry. The dispersion diagram in $x$- direction of the sample unit cell in Figure 2.2 is presented in Figure 2.4.

2.3.2 Effective Refractive Index

For a free space wave with frequency $\omega_0$, the wave number $k_0$ is:

$$k_0 = \frac{2\pi}{\lambda_0} = \frac{\omega_0}{c} \quad (2.5)$$

For surface waves impinging on a periodic structure, wave propagation cannot be investigated with plane wave response, but with dispersion relation of this surface. For a 2D-periodic structure, in a full period of $d_x = d_y = p$, the phase shift is $2\pi$. Hence inside one period, from $\Gamma$ to $X$, the $x$- direction phase variation is from 0 to $\pi$, while $y$- direction phase stays 0.

$$\begin{align*}
    phaseX &= 0 \text{ to } \pi \\
    phaseY &= 0
\end{align*} \quad (2.6)$$
Thus, wave number \( k \) (also the propagation constant \( \beta \)) follows a frequency dispersion relation in Equation 2.7. In CST, the phase shift is changed along the irreducible Brillouin zone boundary and frequencies of eigenmodes are obtained accordingly.

\[
k(\omega) \cdot p = 0 \text{ to } \pi \tag{2.7}
\]

At different frequencies, the effective refractive indices \( n_{\text{eff}} \) are obtained by the ratio of \( k \) and \( k_0 \). An example of 20 GHz frequency wave is marked by navy dashed lines in Figure 2.4.

\[
n_{\text{eff}} = \frac{c}{v} = \frac{\omega/k_0}{\omega/k} = \frac{k}{k_0} \tag{2.8}
\]
Chapter 3

Unit Cell Design

There are three main considerations in designing a proper unit cell: 1. **Sufficient refractive index for a Luneburg lens.** That means the unit cell must provide up to $\sqrt{2}$ effective refractive index. 2. **Small frequency dispersion.** We would like a wide band antenna, hence the $n_{\text{eff}}$ versus frequency curve should be as flat as possible. 3. As **wide air gap between the two layers as possible.** A tiny gap is very hard to keep stable, while a larger gap could facilitate the later required flare design since the impedance difference between the gap and free space would be smaller.

### 3.1 Holey Structure

In order to achieve the easiest manufacturing process, the first design attempt is a completely holey unit cell in Figure 3.1. This configuration can be obtained with simple drilling machine tools. By the definition of glide symmetry, the unit cell contains two off-shifted layers. A short-circuited hole is in the bottom layer, and the top layer is shifted half period away from bottom in both $x$- and $y$- directions.

![Diagram](image)

(a) two off-shifted layers

(b) top view of the bottom layer

Figure 3.1: Holey unit cell configuration.
There are four parameters under optimization in this model, to make the lens operate well in required frequency band: periodicity $p$, layer thickness $t$ (which limits the maximum height of hole $h$), hole diameter $D$, and gap thickness $g$ between the two layers. After determining the optimal parameters above, the height of hole $h$ will be varied to tune the effective refractive index, as is shown in Figure 3.2. It is clear in the figure that $n_{\text{eff}}$ did not reach 1.4 for a Luneburg lens application. The complete design procedures are given in following subsections.

### 3.1.1 Deciding Air Gap Thickness $g$

Based on the theory in Chapter 2, we see how the $n_{\text{eff}}$ gradually changes with frequency in Figure 3.2. At different depth of hole, the periodic structure will have different dispersion relation, hence more than one curves are presented. With proper interpolation, the relation between $n_{\text{eff}}$ and $h$ can also be plotted at 28 GHz for different $g$, as is shown in Figure 3.3.

Applying control variable method, there are some presupposed parameter values:

$$D = 2.92 \text{ mm} \quad p = 3.2 \text{ mm} \quad t = 1 \text{ mm}$$

We can tell from Figure 3.3 that the smallest air gap $g = 0.2 \text{ mm}$ results in the highest $n_{\text{eff}}$, however it is far below $\sqrt{2}$. To see the potential of this structure, we
picked \( g = 0.2 \) mm and increased hole height \( h \) as well as layer thickness \( t \), looking forward to achieving a higher \( n_{\text{eff}} \), though this turned out to be insufficient either.

3.1.2 Deciding Layer Thickness \( t \)

Now we keep the other parameters constant:

\[
D = 2.92 \text{ mm} \quad p = 3.2 \text{ mm} \quad g = 0.2 \text{ mm}
\]

and we increase the layer thickness to \( t = 2 \) mm, and the hole height \( h \) up to 1.8 mm. The relation between effective refractive index and hole height is presented in Figure 3.4.

As illustrated in Figure 3.4, the increment of \( n_{\text{eff}} \) with \( h \) turns flat when \( h \) is large. The reason is that the holes are waveguides below cut-off for 28 GHz wave. Waves can not penetrate or propagate in these holes but just oscillating close to the opening. Thus \( h \) does not need to exceed 1.8 mm, where the dispersion relation starts to be invariant as evanescent waves cannot enter any further. The layer thickness \( t \) does not matter as long as it’s larger than \( h \).

However, the needed value \( \sqrt{2} \) for effective refractive index is still not reached. Moreover, from a practical point of view, if the relation between \( n_{\text{eff}} \) and \( h \) becomes nonlinear, the error control in manufacturing could be worse. But since \( n_{\text{eff}} \) is way too small,
$h = 1.8$ mm is theoretically chosen for later unit cell periodicity and hole diameter optimization.

### 3.1.3 Deciding Periodicity $p$ and Hole Diameter $D$

The last two parameters to evaluate are $p$ and $D$. The relation between $n_{\text{eff}}$ and $D/p$ is plotted in Figure 3.5, where each unit cell periodicity $p$ corresponds to a separate dispersion curve. Here all other parameters are kept optimal in order to achieve the highest effective refractive index:

$$g = 0.2\ \text{mm} \quad t = 2\ \text{mm} \quad h = 1.8\ \text{mm}$$

In Figure 3.5, it is clearly shown that $D/p$ has an optimum at about 0.85 for the highest $n_{\text{eff}}$. And with the increase of periodicity $p$, average $n_{\text{eff}}$ also goes up. However, we can see that when $p = 4.3$, the $n_{\text{eff}}$ versus $D/p$ curve only has the latter half. That is because when periodicity increases, the unit cell’s band gap will shift down, and only with larger $D/p$ can modes exist at 28 GHz.

Since our frequency band is from 25.2 GHz to 30.8 GHz, $p$ is even more confined for unit cell’s response at the upper frequency limit. Thus, if we take advantage of highest $n_{\text{eff}}$ value at $D/p = 0.85$, the largest acceptable periodicity $p$ is 3.6.
Figure 3.5: $n_{\text{eff}}$ versus $D/p$ at 28 GHz. $g = 0.2 \text{ mm}$, $t = 2 \text{ mm}$, $h = 1.8 \text{ mm}$

Now we adopt all the optimal parameters:

$$g = 0.2 \text{ mm} \quad t = 2 \text{ mm} \quad p = 3.6 \text{ mm} \quad D = 0.85p = 3.06 \text{ mm}$$

And the result of $n_{\text{eff}}$ versus frequency is already given in Figure 3.2. Apparently, even with the deepest hole, we cannot achieve sufficient effective refractive index. This implies that the simplest holey configuration is not applicable under the current assumptions on manufacturability and tolerances. However, it would be interesting to see the theoretical outcome if we decrease the gap even more.

### 3.1.4 Theoretically Further Exploring

The only parameter we can play with is the air gap $g$ now. In Figure 3.3, the negative correlation between $n_{\text{eff}}$ and $g$ is presented. Hence we could decrease the gap even more, and keep other parameters optimal to see if a higher effective refractive index is achievable. However, it is important to keep in mind that a smaller gap theoretically involves higher $n_{\text{eff}}$, but it is not a feasible solution due to difficult manufacturing.

The result is given in Figure 3.6. When $g$ is smaller than $0.07 \text{ mm}$, the highest effective refractive index exceeds $\sqrt{2}$, satisfying the Luneburg lens’s requirements. The potential of a holey unit cell has been thoroughly explored in this chapter. Although
it is the simplest configuration, it is not appropriate for our antenna. But if we compare unit cells with and without glide-symmetry, this glide-symmetric holey structure shows very small dispersion that definitely can be utilized in some other applications.

### 3.1.5 Comparison with Ordinary Unit Cells

In Figure 3.7, there are two metasurface configurations without glide symmetry. Both of them have the same bottom layer as the selected one for Figure 3.2, but the left one has a smooth metal top cover and the right one consists of a mirror symmetry. The gap \( g \) is kept as 0.2 mm in all configurations. A dispersion comparison between the two ordinary unit cells and glide-symmetric unit cell is given in Figure 3.8. As we can observe, the glide symmetry technique has a huge advantage in terms of frequency dependence, which can be made use of in ultra-wide band applications.

### 3.2 Hole with One Pin

Since a completely holey structure cannot provide enough effective refractive index, one pin is added inside the hole on bottom layer. The pin works as loading to increase the effective dielectric constant of the unit cell. It keeps waves hovering at the top of the hole for a longer time, thus increasing the effective dielectric constant. The structure is shown in Figure 3.7. Here the hole and pin are both square, but inner
Figure 3.7: Unit cell configurations without glide symmetry.

Figure 3.8: $n_{\text{eff}}$ versus frequency for different unit cell configurations.
corners of the short-circuited hole are rounded because we cannot make sharp corners with cylindrical milling cutters.

The design follows the same procedure as presented in Chapter 3.1. The parameters periodicity $p$, hole diameter $D$, layer thickness $t$, air gap height $g$ as well as milling cutter diameter $dr$ (equivalent to deciding the pin’s side length) are optimized one by one under the same method. And $h$ is still the variable to continuously tune effective refractive index. The detailed steps will not be given again, but some results are directly shown below.

### 3.2.1 Feasible Solutions

Solution 1:

$$p = 3.2 \text{ mm} \quad D = 2.9 \text{ mm} \quad t = 2 \text{ mm} \quad g = 0.3 \text{ mm} \quad dr = 1 \text{ mm}$$

The relation between effective refractive index and frequency is given in Figure 3.10. As is shown, $n_{\text{eff}}$ is high enough for Luneburg lens application, and $h$ does not need to be too large, thus avoids greater dispersions. However, if we are keen to have a larger gap, re-optimization can be made as a second solution:

$$p = 3.2 \text{ mm} \quad D = 2.9 \text{ mm} \quad t = 2 \text{ mm} \quad g = 0.5 \text{ mm} \quad dr = 0.8 \text{ mm}$$

The result of solution 2 is presented in Figure 3.11. Larger $h$ is needed in order to reach required value for $n_{\text{eff}}$, and dispersion somewhat increased. But at the moment, it’s still a feasible solution as the gap $g$ is as large as 0.5 mm. The curves of $n_{\text{eff}}$ versus frequency stopped at some point is due to the band gap of metasurface.
Figure 3.10: $n_{\text{eff}}$ versus frequency for different height of hole. Solution 1.

Figure 3.11: $n_{\text{eff}}$ versus frequency for different height of hole. Solution 2.
3.2.2 Manufacturing Considerations

One consideration now is the wall thickness $wall = p - D$ in the unit cell. In both configurations, wall thickness is only 0.3 mm, which might be a problem for milling. In this case, we wish to adjust $p$ and $D$ to achieve $wall = 0.5$ mm. Yet investigation shows that, $n_{eff}$ has positive relation with $D/p = (p - wall)/p$, as is shown in Figure 3.12. That means, if we don’t want to decrease $n_{eff}$, $p$ has to be increased when $wall$ increases to 0.5 mm. However, as is shown in Chapter 3.1.3, the responding frequency band shifts down when $p$ increases, which means increasing $p$ will lead to more dispersive behavior in our needed frequency range, or even no response at all. Even if we choose to sacrifice $n_{eff}$ to keep a smaller $p$, higher hole depth $h$ will be needed to maintain sufficient $n_{eff}$. And that will also lead to more dispersion. The simulation of Solution 2 with modified wall thickness $p = 3.3$, $D = 2.8$ was done, and the same dispersion evaluation as in Chapter 3.4 was also performed, but the result is very unsatisfying.

In this dilemma, we didn’t adopt a thicker wall solution but kept better electromagnetic property. Luckily, in our pre-production discussion, mechanical engineers thought current dimensions are still workable.

Another manufacturing consideration is the inner corners of the square hole. As is
shown in Figure 3.13 (a), initially we gave the cutter (purple circle) a track in dashed lines. After consulting mechanical engineers, we got the knowledge that the cutter should not standstill at corners before changing feed direction, otherwise there may exists some overcut. Thus we let the cutter follow tangential red arcs at four corners instead. The radius of the curvature is half that of the milling cutter, hence the radii of rounded unit cell corners are $0.75dr$. A 3D illustration of milling cutter’s motion at a corner is presented in Figure 3.13 (b). All simulation results in Chapter 3.2 are with this smooth processing path.

![Milling illustration](image)

Figure 3.13: Milling illustration.

### 3.3 Hole with Four Pins

There are two other possible configurations with even more pins. One is square hole with four square pins and the other has round hole and pins. With more pins, the effective dielectric constant will get enhanced, thus we can have a larger air gap $g$. But processing complexity will also increase with the growth of pin quantity.

#### 3.3.1 Round Pins

The configuration of a 4-round-pin unit cell is presented in Figure 3.14. This structure was first suggested in [16] for 60 GHz application.

As always, all parameters are optimized and hole height $h$ is still the variable to tune effective refractive index. The unit cell is symmetric about both $xz$- and $yz$- planes, giving rise to isotropy in the $xy$- plane where periodicity exists. Here the ”isotropy”
neglects the tiny difference between $0^\circ$ and $45^\circ$ propagation direction as explained in Chapter 2.3, and this also applies to all other unit cells discussed in Chapter 3. The dimensions are:

\[ p = 3.2 \text{ mm} \quad D = 2.9 \text{ mm} \quad t = 2 \text{ mm} \quad g = 0.4 \text{ mm} \quad pd = 0.7 \text{ mm} \quad c = 0.6 \text{ mm} \]

In this case, under simple calculation, the diameter of milling cutter $dr$ should not exceed $0.18 \text{ mm}$, which could be a problem for practical manufacturing. Nevertheless, the electromagnetic property of the proposed unit cell is shown in Figure 3.15.

### 3.3.2 Square Pins

If we modify the 4-round-pin configuration to square, we could possibly have larger milling cutter diameter, and easier milling track. The new design is given in Figure 3.16. It has very similar structure as the 4-round-pin unit cell, but with different dimensions and electromagnetic properties. The new dimensions are:

\[ p = 3.2 \text{ mm} \quad D = 2.9 \text{ mm} \quad t = 2 \text{ mm} \quad g = 0.5 \text{ mm} \quad pd = 0.7 \text{ mm} \quad c = 0.5 \text{ mm} \]

Hence the milling cutter diameter $dr$ could be 0.5 mm wide, making it more practical than the former 0.18 mm limit. Also the processing track should be simpler than that of the round configuration.

The performance of the 4-square-pin unit cell is shown in Figure 3.17. Comparing Figure 3.15 and Figure 3.17, we could conclude other advantages of the square solution. First, the general effective refractive index is higher, which means we could have
a larger gap $g$. Second, the relatively small required hole depth $h$ decreases dispersion. Third, the increment of $n_{\text{eff}}$ versus $h$ is more uniform, giving us less problems related to manufacturing tolerances.

Since the thesis is not about electromagnetic wave propagation theory, we did not go through the theoretical electromagnetic principles behind these differences. It is supposed that the difference is due to a larger hole in the squared-pin configuration. More energy thus can enter the hole and oscillate at the top, leading to higher effective dielectric constant. Detailed and more accurate explanation could be given after further exploration.

### 3.4 Final Unit Cell Choice

All in all, only one unit cell configuration can be applied in our Luneburg lens design. It is already clear that the completely holey structure does not satisfy our requirement, and the 4-round-pin unit cell is inferior to the 4-square-pin one. Hence the final choice would be made among the two solutions of hole-with-one-pin configuration and hole-with-four-pin (square) structure.

Though the electromagnetic performance of the 4-square-pin unit cell is quite good in simulation, from a practical point of view, we would choose the single-pin configura-
Figure 3.16: Hole-with-four-pin (square) unit cell configuration.

Since its fabrication is easier and more economical. The two single-pin solutions’ parameters are listed again here:

Solution 1: $p = 3.2$ mm $\quad D = 2.9$ mm $\quad t = 2$ mm $\quad g = 0.3$ mm $\quad dr = 1$ mm
Solution 2: $p = 3.2$ mm $\quad D = 2.9$ mm $\quad t = 2$ mm $\quad g = 0.5$ mm $\quad dr = 0.8$ mm

Their effective refractive index versus frequency performances were already given in Figure 3.10 and Figure 3.11. Though Solution 2 has a larger air gap $g$, it also has higher dispersion. As a compromise of simplicity of manufacturing and optimum antenna behavior, we investigated the dispersion of solution 1 based Luneburg lens first, to see if this dispersion level is acceptable.

### 3.4.1 Dispersion Exploration

To generate a metasurface Luneburg lens and observe dispersion may become too time-consuming, hence a smarter way was chosen by creating a dielectric step-index Luneburg lens with dielectric constant of the supposed unit cell.

The dielectric Luneburg lens has 10 layers, with $14\lambda$ diameter, which is $150$ mm at $28$ GHz. It follows the equal permittivity increment rule given in [3], that the relative permittivity decreases from 2.0 to 1.1 from center to border. The $m_{th}(m = 0, 1, \cdots, 9)$ layer has relative permittivity

$$\epsilon_{rm} = 2 - \frac{m}{10} = 2.0, 1.9, 1.8, \cdots, 1.1, \quad (3.1)$$
Figure 3.17: $n_{\text{eff}}$ versus frequency for different height of hole.

Unit cell with four square pins.

and boundaries of $m_{\text{th}}$ layer are where

$$
\epsilon_r = 2 - \frac{m \pm 0.5}{10}.
$$

(3.2)

Since Luneburg lens has a relation between permittivity and radial position

$$
\epsilon_r = n^2 = 2 - \left(\frac{\rho}{R}\right)^2,
$$

(3.3)

where $R$ is the radius of the full Luneburg lens, we could calculate the position $\rho_m$ of $m_{\text{th}}$ layer’s boundaries

$$
\frac{\rho_m}{R} = \sqrt{\frac{m \pm 0.5}{10}}.
$$

(3.4)

Thus the outline of a dielectric Luneburg lens is determined.

To demonstrate dispersion of the lens, $n_{\text{eff}}$ of the unit cell is needed at different frequencies. As is explained, $n_{\text{eff}}$ only depends on hole height $h$. Thus the independent variable $h$ should be calculated first following the Luneburg lens requirement at center frequency 28 GHz, that is, we first derive the relation $n_{\text{eff}} (h)$ at 28 GHz from Figure 3.10 with polynomial fitting, then calculate desired $h$ for given $n_{\text{eff}} = \sqrt{\epsilon_r} = \sqrt{2}, \sqrt{1.9}, \ldots, \sqrt{1.1}$. After that, the unit cell will be simulated again with the obtained set of $h$, so that $n_{\text{eff}}$ and $\epsilon_r$ can be extracted at lower frequency limit 25.5 GHz and higher frequency limit 30.8 GHz, as is shown in Figure 3.19.
Figure 3.18: Luneburg lens with 10 dielectric layers

Figure 3.19: $n_{\text{eff}}$ versus $h$ at three typical frequencies. Solution 1.
Figure 3.20: Far field Gain Abs ($\theta = 90^\circ$) based on unit cell in Solution 1.

Table 3.1: Far field Performance of dielectric Luneburg lens.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>25.2</th>
<th>28</th>
<th>30.8</th>
</tr>
</thead>
<tbody>
<tr>
<td>Main lobe Gain Abs (dB)</td>
<td>11.2</td>
<td>10.9</td>
<td>10</td>
</tr>
<tr>
<td>3dB beam width</td>
<td>3.6°</td>
<td>3.1°</td>
<td>2.9°</td>
</tr>
<tr>
<td>Side lobe level (dB)</td>
<td>−10.5</td>
<td>−11.5</td>
<td>−7.5</td>
</tr>
</tbody>
</table>

A CST Microwave Studio dielectric lens model was created with dielectric constant $\epsilon_r$ set at each frequency. The center of the 2D concentric dielectric lens lies on the origin of coordinates. A discrete port is put at $R = 7\lambda = 75$ mm, hence an equivalent air layer is created outside the dielectric lens. The far field comparison of three typical frequencies (25.2 GHz, 28 GHz, 30.8 GHz) is presented in Figure 3.20. Detailed performances are listed in Table 3.1.

From the comparisons, we see the gain and side lobe level at 30.8 GHz is already not so good, let alone the performance of unit cell in Solution 2. Hence, our choice will be the hole-with-one-pin unit cell in Solution 1, while further optimization can be done in the lens layout design step in order to decrease dispersion at the upper frequency limit.
Chapter 4

Lens Layout

Once the unit cell is decided, the lens layout can be taken into consideration. The main purpose is how to arrange the metasurface lens in a simple and cost-effective way. However, the unit cell’s performance at high frequency is still limited, hence we should first figure out another way to achieve better overall performance.

4.1 Design Frequency Shift

A possible solution is to have a higher design frequency. Since the unit cell’s effective refractive index is more dispersive at high frequencies, as plotted in Figure 3.10, we can assign the perfect response at a frequency higher than 28 GHz. That means, to design the hole height $h$ for $n_{\text{eff}} (h)$ perfectly fitting Luneburg lens’s requirement at a higher frequency $f_c$. Thus the average effective refractive index inside the whole frequency band will be smaller than before, but $n_{\text{eff}}$ will decrease more at higher frequencies since it’s more dispersive there.

The same procedure as in Chapter 3.4.1 is performed to check dispersions, and determine the best design frequency. In Table 4.1, detailed performances of dielectric Luneburg lens based on the selected unit cell are given with different design frequencies. We can plot out them to compare intuitively, as is shown in Figure 4.1.

The major change happens at high frequency limit. It can be seen that when design frequency increases, gain and side lobe level improve continuously at 30.8 GHz. Though the properties deviate a little bit for 25.2 GHz and 28 GHz, overall performance changes for the better. Hence we select 29.5 GHz instead of 28 GHz as the design frequency where our designed Luneburg lens has the best behavior.
Figure 4.1: Dispersion comparison for different center frequencies.
Table 4.1: Far field of dielectric Luneburg lens with different design frequencies.

<table>
<thead>
<tr>
<th>Design frequency (GHz)</th>
<th>Frequency (GHz)</th>
<th>Main lobe Gain Abs (dB)</th>
<th>3dB beam width(°)</th>
<th>Side lobe level (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>28</td>
<td>25.2</td>
<td>11.2</td>
<td>3.6</td>
<td>-10.5</td>
</tr>
<tr>
<td></td>
<td>28</td>
<td>10.9</td>
<td>3.1</td>
<td>-11.5</td>
</tr>
<tr>
<td></td>
<td>30.8</td>
<td>10</td>
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<tr>
<td>28.5</td>
<td>25.2</td>
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<td>3.6</td>
<td>-10</td>
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<tr>
<td></td>
<td>28</td>
<td>10.9</td>
<td>3.1</td>
<td>-11.8</td>
</tr>
<tr>
<td></td>
<td>30.8</td>
<td>10.3</td>
<td>2.9</td>
<td>-8.4</td>
</tr>
<tr>
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<td>25.2</td>
<td>11.5</td>
<td>3.7</td>
<td>-10.3</td>
</tr>
<tr>
<td></td>
<td>28</td>
<td>10.9</td>
<td>3.1</td>
<td>-11.8</td>
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<tr>
<td></td>
<td>30.8</td>
<td>10.5</td>
<td>2.9</td>
<td>-8.9</td>
</tr>
<tr>
<td>29.5</td>
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<td>3.7</td>
<td>-9.8</td>
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<td>28</td>
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<td></td>
<td>30.8</td>
<td>10.7</td>
<td>2.8</td>
<td>-9.6</td>
</tr>
</tbody>
</table>

4.2 Metasurface Lens Design

Since the lens has two metasurface layers in glide symmetry, we would like to figure out a way to manufacture them identically. If we simply shift one layer by half period, the counterpart unit cells in two layers will have different distance to the origin, thus different \( n_{\text{eff}} \) will there be according to Luneburg lens’s formula \( n = \sqrt{2 - (\rho/R)^2} \). To solve this problem, the construction will keep both layers centered 1/4 period away from the origin, as is shown in Figure 4.2.

In Figure 4.2, the blue gridding is the top layer while the red one is the bottom. System coordinate is settled at the origin marked with red and green arrow axes. Both layers are shifted 1/4 period in \( x \)- and \( y \)- directions (marked in \( u \)- and \( v \)- in Figure 4.2), that we can simply flip one layer over a \(-45°\) axis (black dashed line) to obtain the other one. Thus the we only need to manufacture one model twice instead of two different models.

The metasurface model is generated in Ansys Electronic Desktop (HFSS) with HFSS-API (application programming interface) controlled from Matlab. Taking the gap in between two layers as a rectangular aperture, the radius of metasurface Luneburg
lens is decided from a 3 dB beam width formula [29]:

$$\Delta \theta = 51^\circ \frac{\lambda}{\text{aperture width}}.$$  \hspace{1cm} (4.1)

Since our requirement of 3 dB beam width is 5°, the Luneburg lens diameter, equivalent to \textit{aperture width} in Equation 4.1, will be $2R = 10\lambda \approx 107\text{mm}$. With a loop program, 878 unit cells are created in each layer. The appearance of one layer is presented in Figure 4.3, where a discrete port is set at $r = 56\text{ mm}$ instead of $r = R \approx 53\text{ mm}$ because the Luneburg lens radius $R$ measures the distance from outermost unit cell’s center to origin, yet the feeding should be on the outside of outermost unit cell’s edge.
The model is simulated in CST MWS Transient Solver, and we got E-field in the $x$-$y$ plane as shown in Figure 4.4. Spherical waves were emitted from the discrete port, and got transferred into plane wave through the Luneburg lens. However, we cannot see the reflection coefficient or far field now because the boundaries around the gap are perfectly matched layers, that neither reflection will happen at the border nor any far field will be generated. Therefore, the next step is to design a flare in between the antenna aperture and free space, and set boundaries as Open (add space) in CST, to simulate the real conditions of the antenna.
Chapter 5

Flare and Feeding Design

In order for the Luneburg lens antenna to work in real life, two other components need to be designed apart from the metasurface lens. They are a matching flare at the end of lens, and a practical feeding. A flare can greatly decrease the reflection problem at the edge of lens, as the height of aperture could gradually increase thus no sudden impedance variation occurs at the border. A practical feeding is what we can use in real life, as the discrete port we adopted in Chapter 4 is just a physical concept, not a feeding in reality. In the end, the overall performance of the antenna should take all components into consideration, hence all parts should be matched well apart from having good performance on themselves.

5.1 Flare Design

5.1.1 A Sliced Model

The flare contains six sectors. Seven points are set and optimized in CST instead of assigning numerous points following a math curve. Experience proves that this is completely enough for an appropriate flare. To save simulation time, instead of optimizing the complete flare structure on the metasurface lens, we considered a slice of PEC parallel plate with flare as shown in Figure 5.1. The red dots are where the variable points lie, and a waveguide port is set as feeding in parallel plate aperture. The gap height in between parallel plate is the same as the one in the metasurface lens.

First we have PMC boundary condition on two cut-sides of the slice, thus the waves propagating inside parallel plate will be TEM, as the aperture has PMC boundaries on left and right, and PEC on top and bottom. The plane wave will incident normally on the edge of flare, which is one situation the flare will face in the lens antenna. The
Another situation is to have PEC boundary on two sides, then the aperture will become a waveguide. The slice width must be appropriate in this case, thus the designed frequency range will not be below cut-off, and only one mode exists. As is known, the fundamental mode inside a waveguide is $\text{TE}_{10}$, and it will go forward in a zig-zag way since the waveguide wavelength is always longer than that in free space. The formula of waveguide wavelength is given in Equation 5.1, where $\lambda_0$ is free space wavelength, and $\lambda_c$ stands for cut-off wavelength of the waveguide. When the wave arrives at the edge of flare, we have oblique incidence.

$$\lambda_{\text{guide}} = \frac{\lambda_0}{\sqrt{1 - \left(\frac{\lambda_0}{\lambda_c}\right)^2}} \quad (5.1)$$

Taking both situations into account, a flare outside circular parallel plate with feed at edge is equivalently simulated. The seven points are optimized to achieve the best
performance, leading to a flare length $a = 20.91$ mm, flare thickness $b = 3.18$ mm as given in Figure 5.1. In Figure 5.2, the optimal reflection coefficient of both conditions is given. In our needed frequency band marked with green bar, $S_{11}$ of oblique incidence is below $-15$ dB, and for normal incidence it is below $-20$ dB.

### 5.1.2 Flare with Circular Parallel Plate

To make sure the flare works well with our metasurface lens, further exploration was done. We combined the optimized flare with a circular parallel plate of the same dimensions as the designed metasurface Luneburg lens. A waveguide port is placed at the edge of the parallel plate, as is shown in Figure 5.3. The diameter of the circular plate is 116 mm, while the radial length of the flare is 20.91 mm, less than $2\lambda$ at 28 GHz. The gap in between the two layers remains 0.3 mm. The waveguide port width is 8.5 mm, whose cut-off frequency is about 17.6 GHz.

The reflection coefficient is presented in Figure 5.4, with our frequency band marked with green bar. Ripples should be due to small reflections from the flare, since their frequency interval matches the distance from waveguide port to the edge of flare. The reflection coefficient is below $-20$ dB in our frequency range, meeting the requirements perfectly. Electric field in $x$-$y$ plane is also shown in Figure 5.5, where bare fluctuation at the edge of the flare indicates that the flare is working fine.
Figure 5.4: Reflection coefficient of circular parallel plate with flare.

Figure 5.5: $E$-field in $x$-$y$ plane at 28 GHz.
5.1.3 Flare with Metasurface Lens

One last thing we checked is the combination of metasurface Luneburg lens and flare. Since the above investigations already verified the feasibility of the flare, it would be interesting to see how it works along with the lens. The diameter of lens part is 116 mm, same as that of circular parallel plate in last section, but longer than the calculated one (with value equal to 107 mm) in Chapter 4.2. The reason is, first, the diameter covered by the unit cells is longer than calculated diameter of unit cells’ center positions; second, a length of parallel plate is added outside of the lens, in order to keep some adjusting margin for practical feeding. The model is shown in Figure 5.6, and reflection coefficient is given in Figure 5.7. We can see the reflection coefficient is below -17 dB in designed frequency band.

5.2 Feeding Design

The power source for antenna is coaxial feed line, but it could not be implemented with the lens properly. Therefore, we need to design a transition component to joint the probe and the lens. The component includes two parts, a coaxial probe to rectangular waveguide transition section, and a stepped horn to connect the waveguide and lens aperture. Full structure is given in Figure 5.8, and the two parts are designed separately.
Figure 5.7: Reflection coefficient of metasurface lens with flare.

Figure 5.8: Feeding structure, $x$-$y$ cut plane view.
5.2.1 Coaxial Probe to Rectangular Waveguide

Initially we want to use standard coaxial probe to waveguide transition element, like waveguide WR28. However, the relatively large flange requires a long waveguide in front of it to maintain multiple feedings around the lens, that we can test the scanning capability. Hence, a self-designed feeding is finally adopted. The coaxial probe is from Rosenberger company, and the data sheet drawings are illustrated in Figure 5.9. It has no dielectric around the metal pin, so we need to design the thickness of air surrounding the pin to achieve approximately 50 Ω impedance. The formula is

\[
Z_0 = 138 \log \left[ \frac{D_{out}}{D_{in}} \cdot \frac{1}{\sqrt{\epsilon_r}} \right]
\]

(5.2)

where \(D_{in}\) is the diameter of metal probe, and \(D_{out}\) being the diameter of dielectric. Here we have air, hence the relative permittivity \(\epsilon_r\) equals 1. The pin diameter is 1.27 mm as seen from Figure 5.9, hence \(D_{out}\) is calculated to be 2.9 mm to provide 49.5 Ω impedance.

The width of the waveguide is the same as the corresponding one of WR28, 7.112 mm, and the height is half of the corresponding one of WR28, being 1.778 mm. TE_{10} mode cut-off frequency is 21.1 GHz. Two irises are added to match the impedance of coaxial probe and waveguide, as shown in Figure 5.10. The irises, located in the
transverse plane of magnetic field, can be taken as inductive shunt elements across the waveguide, and the inductance is directly related to the size of iris itself. Similarly, if an iris is placed within electric field, it is considered a conductive element. Irises can be susceptible to break down under high power conditions, especially the electric plane ones since they concentrate electric field. Hence in our design, only magnetic plane irises are employed.

There are 6 parameter under optimization. $L$, $K$ are the location of coaxial probe in $x$- and $y$- direction, $H$ is the height of metal pin in $z$- direction. For the two identical irises, there are 3 parameters, length $IrisX$, width $IrisY$ and position $IrisP$. The corners are all smoothed to mimic real product appearance from milling technique. In these parameters, $IrisP$ is preliminarily estimated from impedance Smith Chart.

As shown in Figure 5.10, a waveguide port is set at the interface between waveguide and stepped horn, named Port2. If we observe the impedance Smith Chart of $S_{22}$ and de-embed it for a proper distance, we can get a more concentrated relative impedance curve that could be shifted closer to origin by simply adding some induct-
The comparison of $S_{22}$ impedance views are demonstrated in Figure 5.11. After adding two irises on the estimated iris plane, the impedance curve changed from subgraph (b) to subgraph (c), encircling and being closer to the origin that stands for perfect matching.

The reason why we could achieve a more compact curve just by moving towards load is due to the different rotation angles on Smith Chart at different frequencies. Along a lossless transmission line, the reflection coefficient is

$$\Gamma = \Gamma_L \exp(-2j\beta l) \quad (5.3)$$
where $\Gamma_L$ is the reflection coefficient at the load, and $l$ is the length from the load back to the measuring location. Since phase constant $\beta$ relates to frequency, the change rate will be different throughout given frequency band. The impedance curve rotates faster at high frequencies than at low frequencies. Hence an expanded curve (a) can turn into a more compact one (b), making it easier to match with shunt elements.

### 5.2.2 Stepped Horn Connection

The stepped horn connects the coaxial probe-waveguide component and the meta-surface lens. It is designed with 3 steps as shown in the red frame in Figure 5.12. The rightmost section has height $g$ equal to the gap thickness of the lens, and length $d_1$. The next two sections have height $g + t_2$ and $g + t_2 + t_3$ respectively, and length $d_2$, $d_3$. Waveguide to the left of section 3 is the waveguide designed in Chapter 5.2.1, whose height is 1.778 mm in $z$-direction. All five parameters in red are optimized, aiming to match the waveguide and lens aperture perfectly. After optimization, the total length $d_1 + d_2 + d_3$ of the stepped horn is 10.84 mm. The impedance view of reflection coefficient in Smith Chart is shown in Figure 5.13, where the reflection is very small as the curve embraces the origin tightly.

After assembling the waveguide and stepped horn together, the total reflection coefficient is below -20 dB as shown in Figure 5.14.
Figure 5.13: Stepped horn, S-parameter impedance view.

Figure 5.14: Reflection coefficient of whole feeding.
Chapter 6
Complete Design Results

In this chapter, metasurface lens, flare and feeding are assembled together into a full antenna. The correct feed position is investigated, and multiple feedings are arranged to cover certain scanning angles. The reflection, crosstalk, electric field and radiation pattern are all presented.

6.1 Focal Point Determination

The appearance of integrated antenna is shown in Figure 6.1. The red dot is where the feeding aperture is. Its location is varied from $r = 56$ mm to $r = 58$ mm to determine the best focal point. Far field directivity, side lobe level, and 3 dB beam width are all compared at different frequencies as presented in Figure 6.2.

Figure 6.1: Bottom layer of integrated lens.
Figure 6.2: Far field comparison of different feed position.
Figure 6.3: Electric field in $x$-$y$ plane at 28 GHz, for full lens with one feeding.

Table 6.1: Far field performance of full lens with one feeding.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>25.2</th>
<th>28</th>
<th>30.8</th>
</tr>
</thead>
<tbody>
<tr>
<td>Main lobe Gain Abs (dB)</td>
<td>16.7</td>
<td>18.2</td>
<td>19.2</td>
</tr>
<tr>
<td>3dB beam width</td>
<td>$6.6^\circ$</td>
<td>$5.7^\circ$</td>
<td>$5.0^\circ$</td>
</tr>
<tr>
<td>Side lobe level (dB)</td>
<td>$-14.5$</td>
<td>$-18.5$</td>
<td>$-16.8$</td>
</tr>
</tbody>
</table>

When $r = 57.5$ mm, the directivity and 3dB beam width are both optimized at all three frequencies in a global aspect, while side lobe level is also one of the best choices. Hence the joint feeding will be implemented at $r = 57.5$ mm.

### 6.2 Single Feeding Result

The full antenna was simulated in both CST Transient Solver and Frequency Solver, to double check the results. Electric field and reflection coefficient are presented in Figure 6.3 and 6.4 respectively. The electric field in Figure 6.4 proves the feasibility of the integrated antenna. Near-spherical waves generated in waveguide feeding are transformed into plane waves after the lens and got transmitted properly by the flare. In Figure 6.3, we could see $S_{11}$ in different solvers matches well, and is below -16 dB in the required frequency range. Far field at three typical frequencies are also plotted in Figure 6.5, with detailed data given in Table 6.1. Comparing with Table 1.1 in Chapter 1, the beam width is a little bit larger than specified value $5^\circ$, but still acceptable. All results verified the validity of the design method.
Figure 6.4: Reflection coefficient of full lens with one feeding.

Figure 6.5: Far field Gain Abs ($\theta = 90^\circ$) of full lens with one feeding.
6.3 Multiple Feedings Result

Further exploration is about the scanning capability, that is, the performance of the Luneburg lens antenna with multiple feedings. The required scan range is ±60°, but due to implementation constraints, primarily space required for waveguide feeds, a scan range of ±50° was selected.

The configuration is shown in Figure 6.6, where 11 feedings are arranged from −50° to 50° with 10° separation, and opening sector of the flare is from −55° to 55°. The angles are with respect to y- axis because now the symmetry axis is x- and flipping
axis is \( y \). Port 6 is the central feeding. The beam cross-over level is preliminarily calculated to be around -8 dB. The reason why feedings have different waveguide lengths is a consideration of fabrication. Since coaxial probe in Figure 5.9 is to be connected to every waveguide, we need to leave enough space for its flange and manual screwing. The enlarged view of two adjacent feeds is given in Figure 6.7.

The reflection coefficient of all ports are plotted in Figure 6.8. Solid curves are of short waveguides, and coppery dashed curves are of long waveguides. Clearly same waveguides at different positions have very similar profiles, and all of them are below -16 dB in our frequency band of interest. Typical crosstalk between feedings are also presented in Figure 6.9, where \( S_{2,1} \), \( S_{6,1} \), \( S_{11,1} \) are compared, and \( S_{3,2} \) is also given since adjacent feedings face the highest crosstalk level. We can see that \( S_{2,1} \) and \( S_{3,2} \) are both below -21 dB.

The radiation pattern at three typical frequencies are also illustrated in Figure 6.10. The cross-over level between beams is about -10 dB. 11 ports are excited at three different frequencies, and the degree \( \Phi \) in Figure 6.10 is with respect to \( y \)-axis. Since the metasurface lens is not exactly isotropic, and with the influence of adjacent ports and incomplete flare, every beam has asymmetric side lobes, except the middle beam from Port 6.
Figure 6.9: Crosstalk between typical feedings.
Figure 6.10: Far field Gain Abs ($\theta = 90^\circ$) at different frequencies.
Chapter 7

Tolerance Evaluation

The antenna will be manufactured in-house in Ericsson Kista with Aluminum. Before that, some tolerance evaluation needs to be done for the information of mechanical engineers, and to check the feasibility of this design.

One of the most important factors in fabrication is that the waveguide part will be sealed as a whole, and the cavities will be milled on the bulk of metal. Another crucial aspect is the gap $g$ in between the two layers. We need to make sure it remains stable over the whole lens area, but it might be difficult unless some washers are used.

7.1 Feeding Parameters Investigation

Before performing a parameter investigation, the corners of stepped horn are smoothed under the proposal of the mechanical engineer, since they will also be manufactured through milling. The new appearance is presented in Figure 7.1. No evident electromagnetic change happens, as seen in Figure 7.2.

Figure 7.1: $x$-$y$ cut plane view of stepped horn.
Figure 7.2: $S_{11}$ impedance view comparison of round- and sharp-corner stepped horn.

Figure 7.3: Coaxial probe to waveguide transition structure.
The perfect parameters for the self-designed feeding in Figure 7.3 (same as the one in Chapter 5) are:

\[ L = 2.27 \text{ mm} \quad K = 0 \quad H = 1.22 \text{ mm} \]

\[ IrisX = 0.42 \text{ mm} \quad IrisY = 0.32 \text{ mm} \quad IrisP = 7.45 \text{ mm} \]

Since the coaxial probe needs to be installed manually, its location might not be precisely where it is designed to be. Therefore, a parameter sweep is done for \( L, K \) and \( H \). Each of them are increased and decreased by 0.2 mm, and all other parameters are kept constant during simulation. The reflection coefficient of the integrated feeding accordingly is shown in Figure 7.4. As can be seen, the height \( H \) of probe is the most critical, that requires extraordinarily careful operation. \( L \) and \( K \) have looser requirements.

Another parameter suggested to investigate is the width of waveguide feeding. After a similar parameter sweep procedure, the allowance is concluded to be \((-0.1 \text{ mm},+0.4 \text{ mm})\) to make sure the reflection coefficient remains below -15 dB. The reason why we cannot allow smaller value is due to cut-off of the waveguide. At smaller width, the cut off frequency is too close to our lower frequency limit 25.2 GHz, and the performance deteriorates rapidly.

### 7.2 Flare edge Smoothness Investigation

Since the lens is designed to be stacked in a pile which will allow for vertical steering, the edge of flare is preferably smoothed rather than sharp as in Figure 5.1. The radius of mellow edge is 0.5 mm, and a stacked flare slice is given in Figure 7.5. It contains a middle model, bottom layer of the model above, and top layer of the model below.

The modified reflection coefficient with PEC and PMC boundaries are given in Figure 7.6. It can be seen that though curve shape changed, \( S_{11} \) with PEC boundaries (oblique incidence) is still below -15 dB, and with PMC boundaries (normal incidence), it is below -20 dB.
Figure 7.4: Tolerance Evaluation of coaxial probe’s location.
Figure 7.5: A stacked slice of parallel plate with flare.

Figure 7.6: Comparison of a sliced flare with different edges.
(a) Bottom layer with a washer.  
(b) Electric field in $x$-$y$ plane at 28 GHz.

Figure 7.7: Model demonstration.

Table 7.1: Far field performance of one feeding lens with a washer.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>25.2</th>
<th>28</th>
<th>30.8</th>
</tr>
</thead>
<tbody>
<tr>
<td>Main lobe Gain Abs (dB)</td>
<td>17.5</td>
<td>18.1</td>
<td>18.2</td>
</tr>
<tr>
<td>3dB beam width</td>
<td>6°</td>
<td>5.4°</td>
<td>5.1°</td>
</tr>
<tr>
<td>Side lobe level (dB)</td>
<td>−13.8</td>
<td>−12.1</td>
<td>−11.8</td>
</tr>
</tbody>
</table>

7.3 Gap Height Investigation

The gap $g$ in between the two layers must be kept constant to achieve the best performance. Usually thin foams can be used, but in this design we want no dielectric and the gap is too thin for a robust foam to fit. Hence we could introduce a screw or washer inside the gap, but diffraction will happen and the wavefront might get ruined. Another method is to not use a washer but let the antenna be, and find out at what value of $g$ will the behavior deteriorates.

First, a 2 mm radius Nylon P66 ($\epsilon_r \approx 4$) cylinder is defined at the front side of the gap, as is shown in magenta Figure 7.7 (a). The electric field get distorted to a certain extent as in Figure 7.7 (b), and its far field at three typical frequencies is plotted in Figure 7.8. The detailed radiation patterns are listed in Table 7.1. We can see that the side lobe level is already poor that we cannot accept any larger washer radius.

Another method is to use no upholder, just let the sealed feeding part support the
Figure 7.8: Far field Gain Abs ($\theta = 90^\circ$) with a washer.

Table 7.2: Far field performance of one feeding lens with different $g$.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>$g = 0.26$ mm</th>
<th>$g = 0.34$ mm</th>
</tr>
</thead>
<tbody>
<tr>
<td>25.2</td>
<td>28</td>
<td>30.8</td>
</tr>
<tr>
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<td>30.8</td>
<td>25.2</td>
<td>30.8</td>
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<table>
<thead>
<tr>
<th>Main lobe Gain Abs (dB)</th>
<th>17.5</th>
<th>17.9</th>
<th>17.7</th>
<th>16.2</th>
<th>17.5</th>
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<tr>
<td>3dB beam width</td>
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<td>6.2°</td>
<td>6.7°</td>
<td>7.4°</td>
<td>6.5°</td>
<td>5.6°</td>
</tr>
<tr>
<td>Side lobe level (dB)</td>
<td>−15.5</td>
<td>−16.6</td>
<td>−17.3</td>
<td>−11.5</td>
<td>−12.8</td>
<td>−15.0</td>
</tr>
</tbody>
</table>

two layers. This situation is hard to simulate identically since the bending pattern could be extremely complex, when we have one side fixed and the other side free. Hence the model in CST is adjusted with a uniformly varied gap, and tolerance limit is checked during the variation of $g$.

The largest acceptable change is 0.04 mm after investigation, which means $g$ must be between 0.26 mm and 0.34 mm if it varies uniformly. Though this is not the real situation, we can get the feeling of how thick we can accept and let the real height be in this range although the lens bends complexly. The electric field and reflection coefficient for different values of $g$ are shown in Figure 7.9 and Figure 7.10. And radiation pattern with details are given in Figure 7.11 and Table 7.2 respectively.
Figure 7.9: Electric field in $x$-$y$ plane at 28 GHz.

(a) $g = 0.26$ mm. (b) $g = 0.34$ mm.

Figure 7.10: Reflection coefficient at different values of $g$. 

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Figure 7.11: Far field Gain Abs ($\theta = 90^\circ$) with different $g$.
Chapter 8
Conclusions and Future Work

In this thesis, a low-profile 2D metasurface Luneburg lens antenna is designed. It is all-metal, with two glide-symmetric layers and flares, as well as 11 waveguide feeding ports. The antenna performs well from 25.2 GHz to 30.8 GHz, showing a reflection coefficient smaller than -15 dB, and crosstalk lower than -20 dB. The beam width in the whole frequency range varies from 6.6° to 5°, satisfying the requirement. The tolerances of lens are also estimated for the easiness of manufacturing. It can be seen that the air gap thickness between the two layers is the most critical one and might need further handling. When material is set as lossy Aluminum, whose electric conductivity equals $3.56 \times 10^7$ S/m, the loss in metal is about 12% of accepted power in simulation, corresponding to $10 \log_{10}(1 - 12\%) = -0.555$ dB variance in realized gain, as can be seen from Figure 8.1 and Table 8.1.

The future work includes the fabrication of the whole lens and measuring. We can test the prototype’s performance without a gap upholder since the feeding sector will be sealed as a solid, providing quite steady support. Otherwise we can employ a small washer at the other side of the lens to sustain the lens even more. Further optimization can also be done for potential mass production, such as a better feeding method to provide larger angular coverage and higher beam cross-over level, or a smarter way to keep the whole metasurface lens antenna more stable and decrease

<table>
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<tbody>
<tr>
<td>PEC lens (dB)</td>
<td>17.7</td>
<td>18.7</td>
<td>19.2</td>
</tr>
<tr>
<td>Aluminum lens (dB)</td>
<td>17.2</td>
<td>18.2</td>
<td>18.6</td>
</tr>
</tbody>
</table>
the leakage from feed assembling, or even change the manufacturing method from milling to plastic molding for a better unit cell solution, as there are less need of considering practical processing difficulty.


