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Millimeter Wave On-chip Antennas and Metamaterial in Silicon

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Millimeter Wave On-chip Antennas and Metamaterial in Silicon

DISSERTATION

submitted in partial satisfaction of the requirements
for the degree of

DOCTOR OF PHILOSOPHY

in Electrical Engineering

by

Shiji Pan

Dissertation Committee:
Associate Professor Filippo Capolino, Chair
Professor Payam Heydari
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2014
To my wife for her devoted love

To my parents for their endless love and support
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ABSTRACT OF THE DISSERTATION

Millimeter Wave On-chip Antennas and Metamaterial in Silicon

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Aggressive scaling of (Bi)CMOS technology has enabled the designs of highly integrated transmitting and receiving system in silicon at millimeter waves (mm-waves). At mm-wave frequencies, when the antenna’s form factor is in the order of several millimeters, the integration of antenna on the silicon chip is becoming feasible. The idea of on-chip antenna (OCA) has triggered considerable interest, as it allows the ultimate on-die integration of the entire wireless transceiver, eliminating the need for any off-chip interconnection. However, designing a high gain and high efficiency OCA is very challenging because of the high permittivity and low resistivity of the silicon or SiGe substrate. An OCA with silicon substrate will lose much of its power inside the silicon, thereby resulting in low radiation efficiency, low gain and possible electromagnetic interference with the active circuitry. To avoid that, a shielding ground plane above the silicon is desired and in that way the insulator layer acts as the antenna substrate. However, because of the extremely thin thickness of insulator layer (around 5 µm to 20µm) the antenna bandwidth is extremely narrow and the radiation efficiency is very low in general.

This work explores the ideas of using metamaterial surfaces (metasurfaces) in OCA design. The metasurface is used for OCA design in two different configurations: one is as a...
reflector below the OCA and the other is as an antenna directly. It is found that under the CMOS environment, the artificial magnetic conductor (AMC) property of metasurface as a reflector cannot be guaranteed at the resonance when the thickness of the metasurface becomes very thin. Based on an equivalent model, the threshold condition of AMC appearance and the AMC bandwidth formula is derived. Meanwhile, the design of using metasurface as a leaky wave antenna (LWA) was implemented at 94 GHz in a 0.18 µm BiCMOS. The results show the widest relative impedance bandwidth and are among the highest gain ever achieved for OCA considering an extremely thin substrate thickness, as confirmed by both simulations and measurements. A similar antenna operating at 0.31 THz was fabricated in a 65 nm CMOS technology and the simulation results are shown in this work.

Several novel OCA designs are also presented, including a rectangular cavity backed slot antenna, a substrate integrated waveguide (SIW) slot antenna and an E-shaped patch antenna. Even though these antennas have extremely thin substrates, these designs show improved impedance bandwidth and gain compared to the previous designs in the literature. For the sake of wideband operation, an on-chip bowtie shaped slot antenna is implemented both as a single element and in a phased array configuration in a 0.18 µm BiCMOS technology.

Additional research into the potential for applying metamaterial in CMOS devices is also explored. The novel concept for applying split ring resonator in on-chip slow wave transmission line (SWTL) design in silicon is presented.
Chapter 1 Introduction

1.1 Millimeter Wave System in Silicon and Potential Applications

With the potential to achieve a higher data rate and faster communication speed, millimeter wave frequency band (commonly defined as the frequency range between 30 GHz to 300 GHz) has attracted lots of interest in research and commercial applications. Figure 1.1 shows the attenuation profile of the electromagnetic frequency spectrum from microwaves to visible lights. Low-attenuation atmospheric windows centered at frequencies of 35-, 94-, 140-, and 220-GHz, which provide transmission through obstacles such as clothing, smoke, dust, fog, and clouds. Millimeter wave imaging applications such as concealed weapon detection, airplane navigation in low visibility conditions, and satellite surveillance have been targeted at these frequencies. Moreover, there are many other applications at millimeter waves including the multi-Gbps high speed wireless link at the unlicensed 60 GHz frequency band [1], anti-collision automotive radar operating at 77 GHz [2, 3], the point-to-point wireless backhaul communication at 71-76, 81-86 and 92-95 GHz [4], and the human medicine treatment in the frequency range between 40-70 GHz [5, 6].

The aggressive transistor scaling and advances in (Bi)CMOS fabrication process has driven an unprecedented growth in high speed data rate communication and wireless technology in the last two decades. Compared with compound semiconductor (e.g., III-V semiconductor), silicon-based processes (CMOS and BiCMOS) offer the great advantage in terms of cost and ease of fabrication. Silicon technologies have already been proven as a viable solution for millimeter wave circuits and systems in [7-9].
1.2 Potential of On-chip Antennas

Traditionally, different modules of a wireless system are separately fabricated using different technologies to achieve the best system performance. For example, the baseband circuits are mostly realized in CMOS process while the power amplifiers are implemented in III-V compound semiconductor technologies. Meanwhile, planar antennas, used in personal wireless devices, are mostly fabricated on the printed circuit board (PCB) type of substrate, e.g., Duroid and FR4 as illustrated in Figure 1.2. This type of integration thus involves multiple chip packaging and requires a relatively large area for the whole system. To alleviate this issue, the system-in-package solutions as in [11] have been examined to reduce the system form factor. However, the antenna, usually the largest component of the system, is still developed outside of the package.
At millimeter wave (mm-wave) frequencies, when the antenna’s form factor is in the order of several millimeters, the new integration techniques including antenna in package (AIP) as illustrated in Figure 1.3 and on-chip antenna (OCA) as illustrated in Figure 1.4 are becoming feasible. In [12], a 60 GHz folded dipole antenna built on fused silica substrate was integrated together with a transmitter chipset inside a package. A short interconnection between the chipset and the antenna feed line was realized by flipping the antenna substrate such that the feed line touches the bond pad on the die directly. Antennas designed on package substrates usually show good radiation performance in terms of radiation efficiency and gain. In [13, 14], the possibility of using wire bond as an antenna directly has been investigated. In [15], a 11.5 dBi broadside gain was achieved from a 2×2 small patch array utilizing wirebond as the interconnection between the die and the feed line of the off-chip antenna. Although chip-to-chip interconnections (e.g., wirebond) show low insertion loss at mm-wave frequencies below 100 GHz, the performance of these antennas still suffers from fabrication tolerance including the misalignment and deviations from desired lengths and shapes [16], which could be a possible detrimental factor at the sub-mm-wave frequencies. In addition, the assembly of AIP requires specialized processes which would increase the cost.

Figure 1.2. Solution of antenna off the package.
Meanwhile, aided by the advances in silicon technologies, the OCA has triggered considerable interest, as it allows the ultimate on-die integration of the entire wireless transceiver, eliminating the need for any off-chip interconnection. With antenna fabricated on the same die as the radio frequency circuit front-end, OCAs offer the much smaller feed line interconnection losses, which could be a substantial advantage for higher frequencies, including sub-mm-waves and Terahertz (THz) systems. Additionally, the OCA solution has the advantages of compactness, reliability, reproducibility, and so forth, simply because it avoids the complex processes required by inter-chip and intra-chip connections. OCAs can be fabricated and assembled in a large scale at a low cost in the CMOS monolithic form. OCAs can be utilized in a large variety of applications such as radio frequency identification (RFID) tags [17], radio frequency sensors [18, 19], radio frequency radars [20, 21], digital clock distribution [22], intra-chip and inter-chip data communications [23], and so forth.
Despite the great potential use of OCAs in a wireless system, it is very challenging to design an OCA with high radiation efficiency and wide bandwidth. In Chapter 2, the fundamental causes of the difficulty in OCA design will be discussed. It is concluded that a shielding ground plane at the lower metal layer (e.g., M1) is demanded for a robust OCA design. Without a shielding ground plane at M1, OCA designs suffer from the excitation of surface wave(s) in the silicon substrate, leading to impairments such as losses, coupling, and higher sensitivity of antenna performance on the die’s dimensions. The improved radiation performance are shown from several novel OCA designs including a rectangular cavity backed slot antenna, a substrate integrated waveguide (SIW) slot antenna and an E-shaped patch antenna. Besides, a bowtie-shaped slot antenna with wideband operation performance was demonstrated in a 0.18 μm BiCMOS both as a single element and in an array implementation. The results are verified by both simulation and measurement.

1.3 Prospect of Using Metamaterial at Millimeter Waves in Silicon

Metamaterials are engineered media characterized by electromagnetic constitutive parameters (permittivity $\varepsilon$, permeability $\mu$, refractive index $n$, and impedance $Z$) with anomalous
values that may allow the arise of interesting properties in their interactions with electromagnetic waves. The concept of metamaterial was brought up in [24], where it demonstrates several examples of periodic structures (including split ring resonators, SRRs) showing artificial magnetism, and how a material can be formed on a larger scale using such elements. In the last decades, research has been spread to apply metamaterial concept in a varieties of application, including antennas [25], absorbers [26], superlens [27], cloaking devices [28] and etc. Because of the intrinsic interesting property, research has been conducted to implement the metamaterial concept in silicon-based processes. In [29, 30], the negative refractive index property of metamaterial has been applied in phase shifter design in a CMOS technology. In [31], a low phase-noise, low-power millimeter wave oscillator design based on metamaterial resonators was proposed.

In this work, the possibilities of using metamaterial in OCA design are investigated. Chapter 3 and Chapter 4 detail the designs of OCA inspired by the concept of metamaterial surface (metasurface), which is composed by a metal layers patterned in periodic structures with a ground plane below. The metasurface is used for OCA designs in two different configurations as discussed in Chapter 3 and Chapter 4 respectively:

(i) The metasurface is placed below the OCA (e.g., dipole or folded dipole) as a reflector, by its artificial magnetic conductor (AMC) or high impedance surface (HIS) property at the resonance. In [32, 33], the metasurface has been used as AMC below the OCA to enhance the antenna radiation performance. The ground plane of the metasurfaces in [32, 33] is placed below the lossy silicon, in order to maintain enough thickness and therefore ensure the typical AMC property. However, it is found that when considering a shielding ground plane at the lower metal layer, the AMC property cannot be guaranteed at the resonance due to the extremely thin
thickness of the metasurface. In this work, we discuss about the threshold condition for the AMC occurrence based on an equivalent circuit model of the metasurface. The discussion also leads to how to improve AMC bandwidth.

(ii) The metasurface is used as a radiator directly. In this configuration, the ground plane is placed at one of the lower metal layer in the silicon process and provides a solid shielding from the substrate and other circuitry. The fundamental radiation mechanism and the design procedures of the antenna are elaborated in Chapter 3. The antenna design is implemented at 94 GHz in a 0.18 µm BiCMOS. For the design centered at 94 GHz, the measured results shows −2.5 dBi peak broadside gain with 8 GHz 3-dB gain bandwidth and an impedance bandwidth larger than 10 GHz. In its class of broadside radiating fully on-chip antennas, with a ground plane on the lower metal layer of the process, and without additional fabrication processing, this structure achieves the widest impedance bandwidth at W-band and one of the highest gain and gain bandwidth. A similar design operating at 0.31 THz is also fabricated in a 65 nm CMOS technology.

In Chapter 5, a novel design for on-chip slow wave transmission line (SWTL) in silicon is proposed. Traditionally SWTLs are realized by increasing the capacitance per unit length. Here the approach is based on placing extremely flat split ring resonators under the signal trace of a transmission line (e.g., a coplanar waveguide). Compared to other SWTLs, the one proposed here offers the feasibility to control also the distributed inductance, resulting in the control of both the wavenumber and the characteristic impedance of the guided wave.
Chapter 2  Designs of On-chip Antennas

2.1 Introduction

(Bi)CMOS technology is considered as an excellent platform for highly integrated systems operating at millimeter wave (mm-wave) frequencies in terms of its low cost and weight. Various transceiver designs without integrated antennas have been implemented on single die using CMOS technology at mm-waves [7, 34]. However, to achieve a fully integrated system, it is highly desirable to incorporate the antenna with the front-end circuit. With an integrated antenna, the system size could be shrunk to an unprecedented level. On-chip antennas (OCAs) have been proposed to minimize antenna feed interconnection losses. It is interesting to note that mm-wave OCAs are useful even when they operate at a frequency much higher than the one in the CMOS active components, for example when self-mixing or other circuit techniques are used [35, 36].

Antenna in package (AIP) is another viable solution for an integrated radiating system at millimeter waves. It provides more design flexibility and shows good performance, including high gain and wide bandwidth at 60 GHz [37]. In [38], 10% bandwidth and more than 80% radiation efficiency was achieved. However, OCA solutions need to be explored to realize a fully-integrated radio on single chip and avoid transmission losses due to chip-to-package interconnections.
2.2 Design Difficulties of High Performance On-chip Antennas

OCAs have already been designed based on dipole, folded dipole, slot and inverted F [39-41] with radiation off (orthogonal to) the chip. Figure 2.1 shows the cross section (lateral) view of a 0.18 µm BiCMOS chip environment with six metal layers over the silicon substrate. For most CMOS processes, the thickness (h) between the topmost metal layer (M6, in the process shown in Figure 2.1) and bottom metal layer (M1, in the process shown in Figure 2.1) is usually only around 5~15 µm. Thus, among those designs in [39-41], to avoid a very low radiation efficiency, the bottom metal layer is not used as ground shielding between insulator layers and the silicon substrate. However, this makes the antenna radiate mainly in the low resistivity silicon substrate, still resulting in extremely low antenna gains (i.e., low efficiency) at mm-wave frequencies. For example, the 140 GHz OCA implemented in 65 nm CMOS by [19] has a measured gain of only –25 dB. Besides the conduction and dielectric loss, the guided mode excited in high dielectric permittivity ($\varepsilon_r \approx 12$) silicon substrate [42] is the cause of another efficiency loss. The Yagi-Uda antenna structure in [43], which radiates in the lateral side direction of the chip, also shows low radiation efficiency. For future applications of mm-wave transceivers with OCAs, such as mm-wave imaging and multi-gigabit-per-second short range wireless communications, it is strongly desirable to achieve high efficiency antennas, which would lead to integrated transmitters and receivers with efficiency much higher than the current state of the art.
Another major concern in the integration of OCAs is the electromagnetic interference (EMI) resulting from mutual coupling between the antenna and the high frequency front end circuits. In OCA solutions, the antenna and circuit front end are utmost close to each other compared with other solutions (e.g., antenna in package and antenna out of package). As discussed in [44-46], surface waves would be important wave mechanisms for electrically thick silicon substrates. Due to strong surface waves excited in an electrically thick silicon substrate, coupling between antenna and other chip component will be a potential issue as shown in [47]. The EMI could affect the system significantly due to the substrate coupling [42], which could considerably degrade the proper operation of the integrated antenna and circuits. Therefore, a ground plane at the bottom metal layer (M1) is strongly desired to suppress the mutual coupling between the OCA and front-end circuitry, for reducing electromagnetic interference. Importantly, a ground plane on M1 would also fully shield the antenna from the substrate rendering the antenna performance robust with respect to variations of the die dimensions [48].
Other studies with a full ground plane at the bottom metal layer have only shown poor performance in terms of gain and/or bandwidth. Though it is true that radiation performance can be improved when using not fully on-chip solutions, or antenna in package solutions [12, 49] or even design antenna radiate in endfire direction [50].

In spite of the challenges as mentioned above, we attempt to overcome the difficulties with novel antenna designs. In the next two sections, a few novel OCA designs will be presented, considering the ground plane either (i) at the lowest metal layer of the silicon process or (ii) below the silicon. The proposed designs provide a state of the art broadside radiation performance compared to those in the literature.

2.3 Design of On-chip Antennas with Ground at M1

In this session, three different OCA designs with a ground plane on the lowest metal layer (M1) are presented. Firstly, we introduce an on-chip slot antenna backed with a rectangular cavity in a 0.18 µm BiCMOS technology. The cavity is used to enhance the slot antenna radiation and shield the radiating section from the lossy silicon substrate. Despite the extremely thin thickness of the cavity, constrained by the CMOS process, the antenna has 5 GHz input bandwidth around 140 GHz, 20 GHz gain bandwidth, and –2 dBi maximum broadside gain. Results have been obtained from two different full wave electromagnetic solvers, HFSS (a frequency domain solver based on the finite element method [51]) and CST (a time domain solver based on the finite integration technique [52]), which are in good agreement. Furthermore, it has been shown the cavity backed antenna is insusceptible to the presence of nearby conductors.
Secondly, we present a design of a 140 GHz E-shaped on-chip patch antenna in a 0.18 μm BiCMOS technology, with full ground plane on the bottom metal layer. The E-shaped patch antenna exhibits –2 dBi peak broadside gain and 10 GHz impedance bandwidth, which is the widest impedance bandwidth achieved so far at D-band among OCA designs with a ground plane at the lower metal layer in the literature.

At last, we demonstrate a slot antenna design based on substrate integrated waveguide (SIW). The design is suitable for a long but narrow chip floor plane.

### 2.3.1 Extremely Thin Cavity-backed Slot Antenna at 140 GHz

A rectangular cavity is placed below the slot antenna to shield the wave from penetrating into the lossy substrate and also to block the wave travelling inside the insulator layers. Although the cavity height \( h \) is only 9.6 μm (around 0.009\( \lambda_d \), where \( \lambda_d \) is the wavelength in silicon dioxide inside the cavity), a good antenna gain is achieved. According to simulation results, the antenna performance could be rarely affected by the outer metals used by the circuitry if designed properly, which makes it a good candidate for an array element.

Figure 2.1 illustrates the side (lateral) view of a chip environment used in this paper. The total thickness from top metal layer (M6) and bottom metal layer (M1) including metal layers is around 12 μm. The silicon substrate has the thickness of 275 μm with the relative dielectric constant of 11.9 and resistivity of 12.5 Ω·cm. All realistic values of metal conductivity (3.39×10^7 S/m for M5~M6 and 2.34×10^7 S/m for M1~M4) are considered in the simulations.
Figure 2. Illustrate the proposed configuration for the on-chip cavity-backed slot antenna, which consists of a cavity, slot aperture, and a passivation layer, on top of a silicon substrate. The slot aperture is implemented on the top metal layer (M6). The cavity is formed by the lowest metal layer (M1) connected to the top metal layer (M6) through vias in between. Due to the limit on the minimal gap between adjacent vias, multiple layers of vias are realized to provide a good shielding for the cavity, as shown in Figure 2.2 (a). The antenna is fed from the edge of the cavity by a designed 50 ohm microstrip line with width 14 µm on M6. It should be noted that the transmission line structure in the cavity side mainly functions as a microstrip line instead of a grounded coplanar waveguide (G-CPW) since $h$ is less than half of the gap (16 µm) between the center and side metals at M6. The slot is excited by the microstrip line and also by the back cavity. The cavity, besides interacting with the slot, is also excited by the field of the microstrip.

The cavity is designed to resonate at its TE$_{110}$ mode close to 140 GHz. Electric and magnetic fields inside the cavity have the expression
\[ E_z = E_0 \sin \frac{\pi x}{a} \sin \frac{\pi y}{b} \quad (2.1) \]

\[ H_x = \frac{j\pi E_0}{\omega \mu b} \sin \frac{\pi x}{a} \cos \frac{\pi y}{b}. \quad (2.2) \]

Note that the magnetic field \( H_x \) has opposite signs for \( y \) larger or smaller than \( b/2 \), and this affects the excitation of the slot. The slot is designed and optimized in terms of gain and impedance bandwidth. The width and length of the cavity, the slot width and length, and the slot offset to the center of the cavity are respectively \( a = 1.2 \text{ mm} \) (=0.56 \( \lambda_0 \) at 140 GHz), \( b = 0.6 \text{ mm} \) (=0.28 \( \lambda_0 \)), \( W = 20 \mu \text{m} \), \( L = 1 \text{ mm} \), and \( D = 10 \mu \text{m} \) as in Figure 2.2 (b).

Figure 2.3 (a)(b) show the simulated gain along the E and H-planes at 140 GHz, the broadside gain and the input reflection coefficient of the optimized antenna versus frequency. The operating frequency is 140 GHz and the input 10dB bandwidth is larger than 5 GHz. The maximum gain of the antenna at broadside (\( \theta = 0^\circ \)) is \(-2 \text{ dBi} \) which is significant for an antenna over a ground plane which is only 9.6 \( \mu \text{m} \) thick. The peak simulated gain appears around 140 GHz and the 3dB gain bandwidth is from 136 GHz to 156 GHz, which allows for Gb/s communication data rates at mm-wave frequencies. The extremely low thickness is the major limitation for obtaining larger bandwidth. In Figure 2.3 (c) and (d), the antenna input reflection coefficient and radiation efficiency are plotted versus frequency for different slot lengths. For slot length values \( L \) equal to 0.8 mm and 0.9 mm, much smaller than cavity length \( (a = 1.2 \text{ mm}) \), two resonances could be distinguished. The lower one is around 137 GHz, which is the resonant frequency of \( \text{TE}_{110} \) mode inside rectangular cavity [53], whereas the higher resonance mainly depends on the slot length \( L \). When \( L \) gets larger, the slot resonant frequency approaches the cavity one and a wider bandwidth is achieved. \( L \) is chosen as 1 mm for a tradeoff between
bandwidth and gain at 140 GHz. It should also be noticed that at the cavity resonant frequency, a large amount of surface current is induced inside the inner boundary of cavity which results in high conduction losses, which degrades the antenna radiation efficiency at 137 GHz as shown in Figure 2.3 (c). This is the reason of the reflection coefficient notch around 137 GHz in Figure 2.3 (b). We stress that the extreme flatness \( h = 9.6 \, \mu m \) of the cavity, constrained by the CMOS process, results in high losses. Indeed it can be shown in [53] that the power of the resonant \( TE_{110} \) mode lost in conduction losses is proportional to \( V^2 (\lambda_d / h)^2 \), where \( V = E_0 h \) is the potential difference between the bottom and top metals at the center of the cavity.

It should be noted that, to satisfy the CMOS design rule related to metal density, 2 \( \mu m \times 2 \) \( \mu m \) rectangular dummy holes are periodically bored on the top and bottom surfaces of cavity, namely M6 and M1. Simulation results prove that as long as the holes are electrically small enough (in our case, 2×2 \( \mu m^2 \)), the effect of these holes is negligible.
To understand how the performance is affected by the dimension of antenna parameters, key parametric analysis results are shown in this section. Figure 2.4 shows the broadside gain while varying the slot offset \((D)\) to the center of the cavity in terms of different slot widths \((W)\). The negative values of \(D\) indicate the slot is on the side of the microstrip feeding (at \(y < b/2\) line), whereas ‘0’ offset means the slot is exactly at the center of the cavity. As shown in Figure 2.4, the peak gain is asymmetric when the slot is placed at opposite sides of the cavity. It can be explained by the fact that there are two different slot feeding mechanisms: (i) the direct feeding from microstrip and (ii) the TE\(_{110}\) mode of the cavity. Indeed, the feeding microstrip, besides exciting the slot, also excites the cavity, which in turn excites the slot. The cavity TE\(_{110}\) resonant mode has a magnetic field \(H_x\) field component as in (2.2), which has opposite signs for \(y\) larger
or smaller than $b/2$. Correspondingly, the excited surface current on the inner part of the top side of the cavity has opposite directions at two sides of the cavity.

Vice versa, the slot excitation from the microstrip does not change when placing the slot on either side of the cavity center ($y = b/2$). Therefore, when the slot is placed at $y > b/2$, the radiation is enhanced since the two exciting mechanisms are constructive. Vice versa, when the slot is located at $y < b/2$, the two exciting mechanisms are destructive, which therefore weaken the antenna radiation. The same trend is observed for any slot widths considered.

Figure 2.5 shows how the cavity size, location and size of the slot affect the broadside gain of the antenna at 140 GHz. The results in the figures are obtained by varying two parameters each time while keeping the others as equal to the values mentioned in Sec. II. The final optimized dimension is indicated by small loops. Figure 2.5 (a) shows that the peak gain exists when the cavity size is around 1.2 mm $\times$ 0.6 mm. It should be noted from Figure 2.5 (a) that the gain could be increased to $-1.8$ dB when cavity length is 1.3 mm, but the impedance bandwidth gets narrower in the case when the cavity gets longer. Figure 2.5 (b) and Figure 2.5 (c) shows that the gain reaches its peak while the length and width of the slot are around 1 mm and
20 µm, respectively. These results verify that the slot and cavity is optimized in the 140 GHz frequency band.

![Graphs showing broadside gain at 140 GHz versus cavity width, slot width, and slot length for different dimensions.](image)

Figure 2.5. Broadside gain at 140 GHz versus (a) width of the cavity for different cavity lengths, (b) slot width for different slot lengths and (c) slot length for different slot widths.
The performance of an OCA is affected by the interferences between the antenna and its surroundings, which include the Si substrate, the silicon-dioxide layers and the metallic parts in the circuit section, as for example shown in [42]. To relieve this problem, the cavity is used to shield the wave from travelling through the silicon dioxide layers and substrate. This also renders the antenna performance not sensitive to the thickness and size of the Si substrate. To demonstrate how effectively this cavity backed slot works, an extra metal piece shown in Figure 2.6(a) is added in the top metal layer to somehow represent the metal used for circuits. This metal piece is placed alongside the slot antenna in the H-plane direction and shares the same width as the cavity. Figure 2.6 (a) shows the gain pattern in E and H-planes at 140 GHz with and without 2mm wide extra metal piece.

Comparing the patterns in Figure 2.6 (a), it is observed that, the peak gain direction stays very close to the broadside direction ($\theta = 0^\circ$), with its value steady around –2 dB while the total radiation pattern is altered slightly. Next, the same examination is applied by placing the metal piece in the E-plane of the antenna as shown in Figure 2.6 (b). The simulation results in Figure 2.6 (b) show that (i) the broadside gain does not change significantly, and that (ii) the gain pattern is changed moderately in the E plane, i.e., the direction of peak gain shifts when the area of the extra metal is increased. This suggests placing the circuit part of the chip in the H-plane direction. Anyway, though further studies should focus on this part of the research, these preliminary results seem to show that the EMI between antenna and circuit is significantly low.
Figure 2.6. (a) E and H-plane gain pattern at 140 GHz with and without extra metal area ($d = 2$ mm) alongside the top metal layer of the antenna in the H plane, (b) E and H-plane gain pattern at 140 GHz with and without extra metal area ($T = 2$ mm) alongside the top metal layer of the antenna in the E plane.

2.3.2 Extremely Thin E-shape Patch Antenna at 140 GHz

Rectangular patch antennas are widely used because of their simplicity and compatibility with printed-circuit technology. It is known that one of the disadvantages of patch antenna is potentially small bandwidth compared with other antennas, although this depends significantly on the antenna substrate permittivity and thickness. As shown in Chapter 10 of [54], a
rectangular patch antenna’s bandwidth is proportional to its substrate thickness. For our case, the ground plane of the patch is placed on the lowest metal layer of the process (i.e., M1). Since the antenna substrate is only 0.9 % of $\lambda_d$ (the wavelength inside substrate) according to the stackup in Figure 2.1, a traditional rectangular patch antenna with ground at M1 leads to an extremely narrow impedance bandwidth.

In this work, a design of an E-shaped patch antenna at the BiCMOS top metal layer (M6) is proposed with a bottom ground plane at M1. Figure 2.1 illustrates the side (lateral) view of a chip environment considered in this work. It is shown that two additional notches in the edge of the feeding introduce an additional resonant frequency besides the resonance of patch itself, as was originally shown in [55] at microwaves with a coaxial feed. Choosing the geometry of E-patch such that the two resonance frequencies are close to each other, the input impedance bandwidth is greatly improved without significant degradation in antenna gain performance. Instead of feeding the antenna with a through-substrate probe as in [55, 56], in this proposed design, the patch antenna is simply fed by a 50 ohm microstrip line from the longer side of the patch in Figure 2.7.

Figure 2.7. Top view of E-shaped patch on M6, with ground on M1.
The total thickness from top metal layer (M6) and bottom metal layer (M1), including metal layers M1 and M6 thicknesses, is around 12 µm. The silicon substrate underneath has thickness of 275 µm with dielectric constant of 11.9 and resistivity of 12.5 Ω·cm. The realistic values of metal conductivity (3.4×10^7 S/m for M6 and 2.3×10^7 S/m for M1) are considered in the simulations.

First of all, a rectangular patch width \( a = 510 \) µm and length \( b = 655 \) µm as shown in Figure 2.7 are chosen to resonate \( f_0 = \frac{c}{2(a+2\Delta a)\sqrt{\varepsilon_r}} \) around 140.2 GHz, in which \( c \) is the speed of light in free space and \( \Delta a \) is the fringing extension factor from Chapter 7 in [54]. The ground plane at M1, below the patch, has a dimension of 570 µm ×700 µm. It should be noted that due to the extremely thin antenna substrate (\( h=10 \) µm), the fringing field is confined in the region very close to the patch edges and therefore the size of the ground plane only needs to be a bit larger than that of the patch. The antenna is fed by a 50 ohm microstrip line as shown in Figure 2.7. Considering the reflection coefficient in Figure 2.8, the rectangular patch antenna without notch shows a narrow impedance bandwidth (less than 4 GHz) with a moderate gain (−1.8dBi).

![Figure 2.8. Comparison of gain and S11 versus frequency for patch with and without notches calculated by a full wave simulator based on the finite element method (HFSS).](image-url)
In what follows we show that we can greatly improve the bandwidth by introducing two additional notches in the antenna feeding side (a design generally known as E-patch antenna). The notches are placed symmetrically offset to the center line of the patch which is indicated by curve-dot line in Figure 2.7. Based on the rectangular patch, a parametric sweep of the dimensional parameters including the notch offset to the patch center line $D$, notch width $W$ and notch length $L$ are conducted to achieve a maximum impedance bandwidth without sacrificing the antenna gain. The optimized design has $W = 165$ µm, $L = 35$ µm, and $D = 180$ µm. With these two additional notches, the antenna input bandwidth as shown in Figure 2.8 is improved to 10 GHz, which is more than twice of the antenna without notches. The reason is that an additional resonance is introduced at 147 GHz as it will be discussed later on in this paper. The antenna peak gain is around $-2$ dBi at 140 GHz, which is close to the case when there is no notches, in other words, it is close to that of a traditional rectangular patch. The 3dB gain bandwidth is close to 20 GHz, between 125 GHz to 145 GHz.

The radiation efficiency if the E-patch antenna fully on chip is shown in Figure 2.9 versus frequency, obtained with two full wave simulators: HFSS and CST. All realistic value of material electrical properties listed above are considered in the simulation. The peak radiation efficiency occurs at 140 GHz, with the value of 18%.
Figure 2.9. Radiation efficiency versus frequency for the E-patch antenna, calculated using two full wave simulators: HFSS and CST.

Figure 2.10 shows the results of radiation pattern of the E-patch antenna in the E and H planes simulated with the two full wave simulators: HFSS and CST. These two full wave simulations, based on completely different numerical methods, are in good agreement, cross-validating the results presented in this paper. The peak gain is around $-2\,\text{dBi}$, and it appears at broadside ($\theta = 0^\circ$, with respect to the z-axis in Figure 2.1) in both E and H planes. Figure 2.11 shows the radiation pattern in both E and H planes at 135, 140 and 145 GHz. It could be observed that the radiation pattern of 135 GHz is similar to the one at 140 GHz while the pattern at 145 GHz is degraded by 4dB compared to 140GHz.
The physical reason of the wide band is understood by showing the top surface current on the E-shaped patch in Figure 2.12 and the complex magnitude of transverse electric field (E-field) normalized to the maximum value on a planar surface above the patch (10 µm away) at two different frequencies (140.5 GHz and 147 GHz) in Figure 2.13, corresponding to the two dips in the curve of the reflection coefficient in Figure 2.8. For simplicity, in both Figure 2.12 and Figure 2.13, only half of the patch, the part below the center symmetry line in Figure 2.7, is
shown. At 140.5 GHz, the current flowing on the patch shown in Figure 2.12 (a) does not interact tightly with the notches and the strongest current are located at the non-radiating edge of the antenna (same observation as in [55]). Meanwhile, the field map of the normalized magnitude of transverse E-field in Figure 2.13(a) shows the strongest amplitude along the two radiating edges of the rectangular patch that contribute most to the radiation, as in a standard patch antenna. Both plots confirm that at 140.5 GHz, the two notches play a minor role in the radiation such that the E-patch functions similarly as a rectangular patch. Instead, at the higher resonance frequency of 147 GHz, the surface current shown in Figure 2.12 (b) is influenced strongly by the two notches as the strongest current is flowing around them. The notches are clearly the reason of the new higher resonance frequency. The filed map in Figure 2.13(b) shows that the strongest transverse electric field is around the notch. As mentioned in [55, 57], the resonance frequency depends on the contour of the current flowing around the notches area. The dimensions of the E-patch determine the frequency of these two resonances that need to be properly designed to ensure a wide band operation.

![Surface current distribution](image)

(a) (b)

Figure 2.12. Surface current distribution on the top surface of the patch at (a) 140.5 GHz, and (b) 147 GHz, showing the two fundamental resonant current distributions. Results are based on full wave simulations (HFSS). Only half of the patch is shown for symmetry reasons.
Figure 2.13. Normalized magnitude of transverse E-field on a surface 10 µm above the patch at (a) 140.5 GHz, and (b) 147 GHz, showing the two fundamental resonant current distributions. Results are based on full wave simulations (HFSS). Only half of the patch is shown for symmetry reasons.

Indeed, in order to understand how the input impedance depends on the E-patch geometry, Figure 2.14 shows the reflection coefficient varying the dimensions of $L$ and $W$ in the proximity of the optimized values ($L = 35$ µm and $W = 165$ µm) respectively. With smaller $L$ or $W$, the two
resonances could be clearly discerned with different resonance frequencies. Vice versa, when $L$ or $W$ is larger than the optimized value, the two resonances tend to merge, which results in a narrower input impedance bandwidth compared to the results from optimized dimensions (the red curve in Figure 2.14). This also shows how the chosen dimensions of the proposed E-patch antenna are optimized so as to achieve a wide bandwidth in D-band.

![Graph showing parametric analysis of the E-patch reflection coefficient and the two resonances versus frequency for different $L$ and $W$ values.](image)

Figure 2.14. Parametric analysis of the E-patch reflection coefficient and the two resonances versus frequency for (a) different $L$ and (b) different $W$ values.

The performance of an OCA is affected by the interference between the antenna and its surroundings, which include the electrically dense Si (or SiGe) substrate, the silicon-dioxide layers and the metallic parts in the circuit sections next to the OCA. As shown in [47], the peak
radiation direction of a monopole antenna with a ground below the Si substrate could be tilted 10°–20° due to the nearby inductor or CPW line. Therefore, to relieve this problem, a ground plane at M1 is strongly preferable to shield waves from travelling into the Si substrate, which has very high dielectric constant. To illustrate how effectively the ground plane works, an extra metal piece shown in Figure 2.15(a) is added on the top metal layer only 50 µm away from the patch antenna to somehow represent a possible additional metal used for circuits. This metal plate is placed alongside the patch antenna, intersecting the E-plane, and shares the same length b as the patch. Figure 2.15 (a) shows the gain pattern in E and H-planes at 140 GHz with and without 1 mm wide extra metal piece.

Comparing the patterns in Figure 2.15 (a), it is observed that, the peak gain direction stays very close to the broadside direction and its value is slightly decreased by 1~1.5 dB. Similarly, the analysis is applied by placing the extra metal plate intersecting the H-plane of the antenna as shown in Figure 2.15 (b). The results show a moderate change in the radiation pattern, and the peak gain is still at the broadside and its value is slightly decreased by 1 dB. Compared with the results in [47] when the ground plane are below Si substrate, the effect due to the existence of the nearby metal on the OCA’s radiation pattern is much weaker when the ground is on M1.
Figure 2.15. Analysis of interference of an extra metal plate next to the E-patch on the radiation pattern. Two cases are considered with metal plate in the E-plane (a), and in the H-plane (b). Radiation patterns are shown for various metal widths ($T = 0$ mm corresponds to absence of extra metal plate). Despite the very close distance (50 $\mu$m) the effect of the plate on radiation is minor.
2.3.3 Extremely Thin Substrate Integrated Waveguide Antenna at 140 GHz

Many novel circuits and devices based on substrate integrated waveguide for millimeter wave applications were developed in recent years using different manufacturing processes on various substrate materials. Several papers [58-60] related to substrate integrated waveguide-antennas operating at millimeter frequencies are also published very recently.

The design of a slot antenna backed with an extremely flat waveguide technique is presented. It shows the merits of ease of fabrication, low cost, least interference to the nearby circuit components and also isolated from the lossy silicon substrate. A $\pm1\text{dBi}$ peak gain is achieved by using a single SIW slot antenna in the longitudinal direction with 20% radiation efficiency. The flat waveguide structure is created within a dielectric layer by adding a top metal over the ground plane and caging the structure with rows of plated vias on either side as was done in [61].

As illustrated in Figure 2.16, it looks like a very flat, short ended and dielectric-filled rectangular waveguide, with highly reduced height compared to the "normal" 2:1 width-height ratio. The waveguide is working with its dominant $\text{TE}_{10}$ mode. Since the cutoff frequency of the dominant mode in a rectangular waveguide is independent of the height, the reduced height does not affect the working frequencies of the waveguide but only reduces the impedance of the wave, which strongly affects losses. This waveguide is thin enough that can be implemented in the top metal layers of CMOS technology.
For the on chip antenna design, the slot antenna based on extremely flat waveguide structure shows the following merit compared to others: ease of fabrication, low cost, least interference by the nearby circuit component and also completely confine the energy away from the lossy p-doped silicon. The only loss need to be concern is the conduction loss on the bottom and top copper layer of the SIW and the wave leakage due to the gaps between the vias. The leakage can be negligible as long as the vias are dense enough and the CMOS circuit design environment is able to meet such condition. And the conduction loss, per unit length, in the waveguide without slots can be estimated by

\[ \alpha_c = \frac{R_s}{a'b\beta k\eta} (2b\pi^2 + a^2k^2), \]

(2.3)
where $a$ and $b$ are the size of the waveguide, $R_s$ is the surface resistance of the copper at 140GHz. According to the theory, the slot has to be offset to the center line of the waveguide and only the $y$ component of current in the slot region disturbed by the slot will contribute to the radiation.

Due to the size limit of the chip environment, only single slot antenna based on extremely flat waveguide structure as shown in Figure 2.17 (a) is investigated. The waveguide is associated with a 50 $\Omega$ microstrip line through a tapered microstrip-waveguide planar transition. The parameters, $L_{\text{trans}}$, $L_{\text{offset}}$, $L_{\text{ext}}$ and $L_{\text{slot}}$ as illustrated in Figure 2.17 (a), are optimized to show their effect on the antenna gain and impedance matching respectively. As shown in Figure 2.17 (b), $-1$ dB broadside gain at 140 GHz is achieved by using a single integrated waveguide slot antenna in the longitudinal direction. However, because of the extremely thin thickness of waveguide, the bandwidth of this type of antenna is comparably small (around 3 GHz). The total area of antenna is $0.6\text{mm} \times 2\text{mm}$, which is suitable for a long but narrow chip floor plan.
2.4 Design of On-chip Antenna with Ground Below Silicon

According to the above discussion, a ground plane at the lower metal layer is preferred for OCA design although that will limit the impedance bandwidth of the OCA. However, for a certain application, a wide operation bandwidth is demanded. For example, for imaging application, the bandwidth of the system would decide the amount of power received and indirectly affect the sensitivity of the imaging system. Out of this purpose, a bowtie shaped slot antenna design showing extreme wideband operation, with a ground plane below the silicon, is presented. The antenna has been implemented using a 0.18 µm BiCMOS technology, both as a single element and also in an array. Using full wave simulation software, the input reflection and gain of the bowtie slot antenna under the influence of a close proximity RF probe is investigated. Later, we present measurements of the input reflection coefficient, radiation pattern, and gain versus frequency. As it will be shown, the antenna has a very good input match ($S_{11} < -10$ dB) over the whole W-band frequency range, partly due to the severe loss in the antenna substrate.
2.4.1 Wideband Bowtie Slot Antenna at 94 GHz

Figure 2.18 (a) shows the die micrograph of the fabricated bowtie-shaped slot antenna, whereas Figure 2.1 illustrates the cross-sectional view of the 0.18 µm BiCMOS process used for antenna fabrication. The bowtie slot antenna is placed on the top-most metal layer (i.e., M6). The antenna structure is symmetrical in both $x$ and $y$ directions, except for the feeding. The antenna dimensions, indicated in Figure 2.18 (a), are (all in mm) $L_1 = 1.4$, $L_2 = 0.55$, $L_3 = 0.05$, $W_1 = 0.9$, $W_2 = 0.46$, and $W_3 = 0.05$. With the exclusion of bond pads, the core antenna occupies $1.4 \times 0.9$ mm$^2$ of chip area. A conductive silver adhesive layer is attached below the silicon germanium (SiGe) substrate to act as the antenna ground.

Figure 2.18. (a) Micrograph of the bowtie shape slot antenna fed by a coplanar waveguide, (b) antenna in the corner of a 5mm × 5 mm chip. The bowtie slot and coplanar waveguide are on M6.
The SiGe substrate in the used technology has a thickness of 275 µm, a conductivity of 12.5 S/m, and a relative dielectric constant of 11.9. The silicon dioxide between M6 and the silicon substrate is roughly 10 µm thick with a relative dielectric constant of 4.2. Above M6, there is a passivation layer with a thickness of 0.6 µm and relative dielectric constant of 7. All the layers are included in the full wave simulation carried out with the finite element method by HFSS. To comply with the metal density rules imposed by the foundry, arrayed 5 µm × 5 µm dummy holes (small black holes as illustrated in Figure 2.18 (a)) are placed around the antenna metal structure on M6. Simulation results prove that the effect of these dummy holes can be ignored, since each hole’s size is much smaller than the guided wavelength of the substrate. A 50 Ω coplanar waveguide (CPW) is employed on M6 to feed the slot antenna and is connected to the bond pad. In the simulation, an HFSS lumped port is used to feed the CPW. The signal trace of the CPW has a 24 µm width, and the gap between the signal and each lateral ground plane is 13 µm. At W-band, the parasitic capacitance of the bond pad could slightly degrade the input matching. To solve this issue, a shunt stub is used to improve the impedance matching of bond pads to 50 Ω at the input (visible in right part of Figure 2.18 (a)). It is noteworthy that both the CPW feed and the shunt stub has negligible effect on the antenna gain and radiation pattern as observed by simulation.

The simulated S_{11} and broadside gain versus frequency with the antenna in the corner of a 5 × 5 mm² die as configured in Figure 2.18 (b) are shown in Figure 2.19(c). The shunt stub is not included in the simulation. The simulated input −10dB bandwidth covers the whole W-band frequency range while the broadside gain at 94 GHz is around −2.5 dBi. The radiation efficiency is close to 18%, mainly due to dielectric losses and surface waves. Direct RF probing is adopted in this work to measure the antenna performance. As shown here, strong electromagnetic
interference is induced between probe and antenna because of their vicinity. Nevertheless, we use the probing method shown in Figure 2.20 because for silicon OCA radiating broadside, it is not straightforward to realize a backside feeding scheme, as used in the measurement of in-package antennas [62], or for other mm-wave antennas. For a backside feeding measurement scheme, even if the RF probe is in the near-field range of the antenna, the probe would have minor effect on the antenna radiation. To realize a backside feeding scheme for a silicon based OCA, an off-chip feed with the help of flip-chip or TSV technology [63] is required.
Figure 2.19. (a) Three-dimensional (3-D) view of antenna and metallic probe-tip, 50 µm above the top surface of the chip, (b) 3-D view of antenna and probe-head attached to the probe tip, (c) $S_{11}$ and broadside gain versus frequency, considering and not considering the presence of the probe. Solid line, no probe. Dashed line, with only probe tip in simulation. Dot line, with both probe tip and probe head in simulation.

![Diagram](image)

Figure 2.20. Schematic diagram of the setup for the gain and radiation pattern measurement of the AUT. A calibrated horn antenna is transmitting and the AUT is in the receiving operational mode. The AUT input reflection coefficient is instead measured by connecting a millimeter head extension directly to the probe.

The antenna is fabricated on a $5 \times 5$ mm$^2$ die, and placed at the corner of the chip, as shown in Figure 2.18 (b). Due to the high dielectric constant and thickness of the SiGe substrate, surface wave modes ($TM_0$ and $TE_1$) are excited inside the antenna substrate. Figure 2.21 shows the broadside gain versus frequency for the proposed antenna in the center and in the corner of the die respectively for various different die dimensions. It can be observed by simulation that when the antenna is put at the center of the die, an increasing size of die area would result in a lower broadside gain because (i) the peak radiation direction becomes significantly tilted away from the broadside and (ii) more surface wave losses by means of dielectric losses exist inside
the lossy SiGe substrate. This makes the antenna unsuitable for broadside applications. However, when the antenna is located close to the edge of the chip, it was observed that the antenna radiation pattern and broadside gain are less susceptible to the chip size, as proven by full-wave simulations.

Figure 2.21. Broadside gain versus frequency of the proposed antenna in (a) the center of a die, (b) the corner of the die, with various die dimensions.
To get a reasonable estimation of the scattering effect caused by the probe in the present configuration, a probe-tip-like metallic part is included in the simulation as shown in Figure 2.19(a). According to the probe datasheet (model Cascade Microtech I110-T-GSG)), the copper probe-tip is 2 mm high at a 45-degree inclination with respect to the chip surface. Because of the modeling difficulty here, instead of physically connecting it to the antenna, the probe-tip is ‘floating’, and it is placed 50 µm away from the top surface of the passivation layer of the chip. Furthermore, to estimate the effect of the bulky probe-head, which is also in close-range of the antenna, an additional simulation containing a copper trapezoidal structure with the probe-head dimensions was performed Figure 2.19(b).

Simulation plots in Figure 2.19 (c) compare $S_{11}$ and gain of the antenna with and without the probe tip. Two cases are considered, as discussed above: one indicated by dashed line, the case where only the probe-tip is included in the simulation and the other by dotted line where both the probe-head and probe-tip are included. Although the simulation is based on simplifying assumptions, and without connecting the probe-tip to the antenna as in measurement, the comparison provides certain useful information on the effect the probe has on the antenna performance. It can be observed that with the probe-tip [Figure 2.19 (a)], the gain can be offset by 1-3dB, and that this is mainly due to the variation of the radiation pattern. Meanwhile, $S_{11}$ only slightly changes, mainly because it is strongly influenced by the losses in the SiGe substrate. When including the whole probe-head in the simulation [Figure 2.19 (b)] the radiation pattern is significantly affected, especially in the E-plane, although the broadside gain is similar to the case when including only the probe-tip in the simulation. These variations are expected because the bulky probe-head is located in the antenna proximity. Therefore, during the measurements,
absorbers should be used to cover the probe head in order to partially shield its interaction with the antenna.

The antenna’s input reflection coefficient is measured while probing the antenna with a W-band RF-probe. A short-open-load (SOL) calibration at the end of the probe-tip is conducted before the measurement. Figure 2.20 illustrates the test setup for the OCA gain and radiation pattern measurements. The signal generator provides a signal in the 12-18 GHz frequency range, which is later up-converted to W-band using a ×6 frequency multiplier. The W-band signal is transmitted by an off-the-shelf horn antenna with 24 dBi gain and received by the antenna under test (AUT), 30 cm away from the horn antenna. A rotation arm was built using aluminum rods and fixed in a high resolution rotary table as shown in Figure 2.22. A metallic plate was used to hold the frequency multiplier and the transmitting (TX) horn antenna tightly on an arm. The radiation pattern of the AUT in the receiving mode is measured by rotating the arm in the E- and H-planes. The AUT output is connected to a probe which feeds an Agilent harmonic mixer and spectrum analyzer.

In the measurement, the antenna gain calibration is based on replacing the AUT with another horn antenna with known gain separated by the same distance as that between the horn and the original AUT. The horn antenna’s phase center was considered as the location of the horn. Since the phase center of the horn antenna is frequency dependent, the calibration power is adjusted based on exactly the same distance between the two phase centers. Insertion losses in the probe as well as an additional 0.5 dB transition loss between the probe and on-chip CPW were all considered in the gain calculation. Absorbers were used to cover the surface of surrounding instruments and components during the gain measurement.
Figure 2.22. Setup for the antenna gain and radiation pattern measurement.

Figure 2.23 shows the comparison between the measured and simulated antenna input reflection coefficient. The simulation results are based on the simulation setup explained above where the probe tip is taken into account. To best compare with the measurement, the shunt stub is also included in the simulation. The proposed bowtie slot antenna shows a very wide $-10$ dB measured input return loss bandwidth covering the entire W-band. Such a wide input impedance bandwidth is also contributed to by the significant overall losses, including surface wave, dielectric, and ohmic losses. To validate the measurements, the same test was repeated on three different chip samples and the results are very consistent.
The antenna gain is calculated based on

\[ G_{AUT} = P_{AUT} - P_{horn} + G_{horn} + Loss_p, \]  

(2.4)
in which all the terms given are in dB scale: \( P_{AUT} \) and \( P_{horn} \) are the power received at the spectrum analyzer when using the AUT and a calibrated horn as receiving antennas, respectively. \( G_{horn} \) is the known gain of the calibrated horn antenna. \( Loss_p \) is the total insertion loss including the probe (1.6dB), the shunt stub (0.4dB), and the transition loss between probe tip and the on-chip CPW (0.5 dB), with a total loss of 2.5 dB. In Figure 2.24, the measured broadside gain, indicated by dots, is obtained by averaging the measured data from three distinct gain measurements, and each frequency point has been averaged including the nearest adjacent frequency points, to smooth out the interference effects. Compared with the simulation curve, there is a 0-4 dB deviation (the maximum difference is at 100 GHz), which is acceptable for OCA measurements at 100 GHz [64].

Figure 2.23. Measured and simulated antenna input reflection coefficient for three different on-chip antennas.
As discussed previously, the accuracy of OCA gain measurements could be degraded by the interference between the probe, probing station, and AUT. For our case, since our antenna is put close to the chip edges, the edge field could be more sensitive to the chip surroundings. The interference due to the probe may occur in two ways: (i) The RF probe itself is acting as an antenna such that it receives/transmits power directly from/to the transmitting horn antenna. In [65], it was shown using simulations that a probe may radiate as a dipole at 60 GHz with $-8$ dBi gain. In our case, using the probe directly as a receiving (RX) antenna, we have measured a received power, radiated by the TX horn, 20 dB lower than the power received by the bowtie slot antenna. This means that in the present case we can neglect this interference. (ii) The wave incident on the probe could be reflected (scattered) to the OCA, and vice versa, causing interferences. The comparison between the simulated normalized radiation patterns in E- and H-plane with/without the probe is shown Figure 2.25. The antenna radiation is varied under the effect of probe tip included in simulation when comparing the black and the red curve. It can be observed that due to the pattern’s variation, there is a 1-3 dB increase in the broadside gain. Also, the radiation pattern is changed significantly in the presence of a probe head, because its size is

![Figure 2.24. Measured and simulated broadside gain of the bowtie slot antenna versus frequency.](image-url)
significantly larger than that of the antenna. To alleviate this problem during radiation pattern measurements, very thin absorbers are used to surround the antenna, and also to partly cover the probe-head. It was observed that with careful placement of absorbers, the interference due to these reflections could be significantly reduced, and a smoother radiation pattern was obtained. Because of these issues, even though the simulation includes the probe, the results shown in Figure 2.19 still may not be fully capturing all the interferences in the measurement.
Figure 2.25. (a) E-plane (b) H-plane simulated normalized radiation pattern of the proposed antenna with and without probe.

The normalized measured radiation pattern of the bowtie slot antenna at 94 GHz is shown in Figure 2.26. The measured pattern is compared against a simulated one, with the probe tip near the AUT. Due to the presence of certain obstacles in the probe station as shown in Figure 2.22, the pattern measurement cannot be measured in the whole E- and H-planes. In the E-plane, the measurement could be done in only one quadrant and the range is limited by 40 degrees from the broadside direction. In the H-plane, the limit is 25 degrees in one quadrant and 55 degrees in the other quadrant of the upper hemisphere. The pattern measurement was repeated twice in each plane and the curves shown in Figure 2.26 are based on averaging these two measured results. The measured and simulated patterns are shown to have similar trends. The ripple of the radiation pattern is probably due to the interference between the probe and the AUT during measurement, and shows the range of inaccuracy when measuring gain at W-band frequencies on
a probe station. Interference can be constructive or destructive depending on the observation angle, and is larger in the E-plane.

![Simulated and measured radiation patterns](image)

Figure 2.26. Measured and simulated, E- and H-plane, normalized radiation patterns at 94 GHz.

### 2.4.2 Array Implementation of Bowtie Slot Antenna at 94 GHz

With a bowtie shaped slot antenna, a highly integrated 9-element imaging receiver comprising a 2×2 super-pixel array with spatial overlapping super-pixels was designed and fabricated in a 0.18 μm BiCMOS process. A unique overlapping super-pixel concept was demonstrated, which employed mechanisms for 7-step amplitude and 9-step delay control settings [66, 67].
A traditional focal plane array is implemented as an $M \times N$ array of elements, with each element feeding one detector, representing one pixel of an image. Image acquisition time for systems employing focal plane arrays is determined by several factors, including the array size, the pixel spacing, the focusing system and the desired spatial resolution of the object. If the object cannot be imaged with one scan, then decreasing scan time requires one or more of the following actions such as, increased array size, reduced pixel spacing, and changes in the focusing system.

A $2 \times 2$ array of elements feeding one detector and constituting one pixel is depicted in Figure 2.27(a). If wideband delay and amplitude control are incorporated in each RF-path from element to detector, then this $2 \times 2$ array becomes a super-pixel, with delay and gain weightings for each RF path. These weightings enable the super-pixel to constrict the field of view of the feed antenna so that more energy illuminates the lens and less is lost to spillover, with a resulting SNR improvement dependent upon the original spillover losses.

In this work, a new system architecture shown in Figure 2.27 (b) is presented [66, 67]. The superpixels overlap and share neighboring elements, and hence, are referred to as spatial-overlapping super-pixels [66, 67]. Each detector is still fed by four elements as in a conventional array of superpixels, but now each element feeds up to four detectors depending on the element’s location within the array. Also, the phase and gain control among the four elements fed to the same detector enables each super-pixel as a $2 \times 2$ phased array.
For successful beam steering, the gain and phase of each RF path from element to detector must be wideband and adjustable. An adjustable wideband true time delay (TTD) circuit and a variable gain amplifier (VGA) in each RF path are provided. The 1:4 power splitter is included so that each element may feed up to four power detectors, and the 4:1 power combiner is used so that each power detector is fed by four elements. The entire RF path before the detector is constructed as a ground-shielded coplanar waveguide (CPW). The TTD circuit has been incorporated into the splitter, providing a significant area savings. A single super-pixel is therefore constructed by combining four RF paths to feed one detector [66, 67].

The proposed 3×3-element array was fabricated in a 0.18 µm SiGe BiCMOS process and a die photo is shown in Figure 2.28. The nine on-chip antennas are placed as one 3×3 array with a separation distance of 1.8mm, or 0.54 λ at 90 GHz. One of the major concerns of using silicon-based on-chip antennas [47] is the risk of interference between antennas and the on-chip circuitry, especially the passive structures (i.e., inductors). This concern is based on the assumption that the antenna is using the silicon as a substrate; however, the situation is different when the antenna is fully backed with a ground plane at M1 and uses the interlayer dielectric as the
antenna substrate. Moreover, for phased arrays, the mutual coupling between antenna elements could be destructive to the radiation pattern. In this design, the low noise amplifier (LNA) provides high gain (~40 dB), and the antenna’s unwanted mutual coupling could adversely modulate the received signal and overwhelm (saturate) the LNA’s output. To reduce these harmful effects, a shielding structure at M1 has been placed beneath the antenna, as explained below, with simulation showing negligible electromagnetic (EM) coupling with the LNAs.

Figure 2.28. Die photo of the 9-element imaging array, each array element with a bowtie slot antenna [66, 67].

Figure 2.29 (a) shows the micrograph of the fabricated antenna. Based on a bowtie-shape slot antenna, the on-chip antenna is built on the top metal layer of the process (M6). Beneath the
slot antenna, there is a shielding structure composed of the ground plane on the bottom metal layer (M1) and the vias between M1 and M6. Figure 2.29 (b) shows the 3-D view of the proposed antenna. The ground plane on M1 is patterned with slots, formed in the same shape and with similar dimensions as the bowtie on M6. The whole antenna is fed by a 50 \( \Omega \) CPW line. A metal stub is added in the middle of the bowtie slot to provide better antenna input impedance matching.

Figure 2.29. (a) Photo of the bowtie slot antenna, (b) 3-D view of the antenna.
It should be mentioned that in spite of the existence of the M1 shielding layer, electromagnetic waves still partially penetrate the shielding, propagate, and establish surface wave modes inside the silicon. To achieve high antenna gain, the silicon substrate was background to a specific thickness (220 μm), which is close to one-quarter wavelength at 90 GHz in silicon. The quarter-wave reflection from the ground plane improves the radiation efficiency.

Figure 2.30 shows the mutual coupling between the center antenna and one of its adjacent antennas in both the E- and H-planes. A comparison between the cases with and without shielding is provided. It can be observed that without the additional shielding structure, the mutual coupling is as high as –22 dB in the E-plane and –28 dB in the H-plane. Meanwhile, with the shielding structure, the coupling between two antennas is 6 dB lower for both E- and H-plane cases.

![Figure 2.30. Simulated antenna mutual coupling in E- and H-plane with and without the shielding layer at M1.](image)
Figure 2.31 (a) shows the layout geometry for the 9 antenna elements of this work. Figure 2.31 (b) shows the simulated broadside gain versus frequency for the four elements inside one subarray, where antenna 1 indicates the antenna closest to the center of the die, antennas 2 and 3 indicate the antennas adjacent to antenna 1 on the H- and E-plane, respectively, and antenna 4 indicates the antenna at the corner of the chip. It can be observed that at 90 GHz, the four antennas show a very similar gain, which is close to 0 dBi. Additionally, over the bandwidth from 80 GHz to 100 GHz, there is at most 4 dB gain variation between the four antennas. Note that the literature show examples of 5 dB gain variation over different antennas [68], where a phased array was designed on a low temperature co-fired ceramic (LTCC) substrate at 60 GHz. For the proposed design, it should be noted that antenna gain variation can be fully equalized by adjusting the gain setting of the VGA, as long as the variation is systematic and predictable.
Figure 2.31. (a) Antenna layout details, (b) simulated broadside gain versus frequency for a 2×2 subarray [66, 67].

It is also noteworthy that antenna 1 and antenna 2 share similar gain over frequency patterns while the same phenomenon happens between antenna 3 and antenna 4. This means that the gain of the proposed antenna is more susceptible to the substrate dimension in the E-plane than that in the H-plane. This implies that if the array is implemented in one dimension using the proposed antenna, it is better to align the array along the H-plane such that each antenna provides similar gain regardless of its location on the die. Although the antennas exhibit gain variation, the simulated reflection coefficients of the four antennas (Figure 2.32) are very similar. Reflection coefficients show good matching from 80 to 104 GHz. Each bow-tie slot antenna occupies 0.5mm² of chip area.
Figure 2.32. Simulated return loss of the four antennas labeled in Figure 2.31(a) [66, 67].

The antenna measurement test setup is shown in Figure 2.33. The test system consists of a laptop, a test fixture, and a thermal control mechanism used to remove any excess heat generated by the receiver chip’s DC power consumption. The test fixture comprises two print circuit boards (PCBs). The first PCB is a custom 2-layer board. It contains an open-top socket that attaches the receiver chip to the PCB, a serial-to-parallel interface that allows communication between the receiver chip and the DAQ/GPIO board, and adjustable voltage regulators that controls the various supply voltages used by the receiver chip and the interface circuitry. The receiver chip is wire-bonded to a 160-pin open-top thin quad flat pack (TQFP), which is placed in the open-top socket. This allows for the testing of multiple receiver chips with a single test fixture, as receiver chips can easily be swapped in and out of the socket. The second PCB is an off-the-shelf DAQ/GPIO board, which communicates with the PC through a USB connection. Custom software provides a simple user interface to reset the receiver chip, to set and check all receiver chip registers, to initiate data acquisition, and to record all data to the
computer. The test fixture was mounted to a rod that can be rotated ±180° with respect to an incident W-band signal 50cm away, and it could be rotated 90° with respect to the rod.

![Test setup for antenna measurement](image)

Figure 2.33. Test setup for antenna measurement [66, 67].

Power calibration was performed prior to testing. The transmit power and frequency were varied on the signal generator, and the output power delivered to the transmit horn was recorded for each transmit power and frequency setting. Next, a horn antenna connected to a power meter was placed 50 cm away from the transmit horn antenna. The transmit power and frequency were again varied and the received power was recorded for each power and frequency setting. The propagation loss in air was calculated to be 65.5 dB for a distance of 50 cm at 90 GHz.

Calibration was required for all receiver measurements due to variation among the RF paths. This variation is mainly comprised of 1) the gain variation among antennas, 2) amplifier gain variation due to process variation, and 3) variation in path loss through the active power splitter, true time delay and power combiner. Generally, the calibration for each RF path is conducted based on the output of its power detector for the ON and OFF states of the transmitter, with the transmit frequency set to 90 GHz, the integration time set to 20 ms and the transmit
power set to a level to prevent saturation of the integrator. The measurement starts with turning on a single RF path, only. The gain of the VGA of this path is swept across all settings when the chip is mounted to receive in the broadside direction. This measurement is repeated for the RF path associated with an adjacent element in the same reference plane. The VGA settings for the two paths are selected so as to provide the same gain in each RF path. Then, with the two adjacent paths of two adjacent elements in the same reference plane turned on together, the radiation patterns are measured under a full sweep of TTD settings combined with the tentative VGA settings. With the transmit horn position fixed, the receiver test fixture is rotated mechanically in 1° steps from −60° to 60° to measure the radiation pattern. At each step, the test program sweeps through all combinations of VGA and TTD settings to eventually generate hundreds of radiation patterns. Based on the measured patterns, the optimal VGA and TTD settings are picked in terms of 1) the peak gain direction, and 2) sustaining similar power levels for different steering angles for each targeted steering angle from −18° to 18° with a step of 4.5°. The setting for each super-pixel is obtained by combining the optimal setting for each 2-element pair in the same reference plane.

A super-pixel is used for antenna beam steering measurements to demonstrate the wideband delay and amplitude control of the super-pixel. The super-pixel was programmed for beam angles from −18° to 18° in 4.5 steps. For each setting the transmit power and frequency remained constant, the receiver-to-transmitter angle was varied from −60° to 60° in 1° increments and the received power was measured. A single RF path was then turned on and the test was repeated. Figure 2.34 shows super-pixel beam steering radiation patterns compared with a single antenna radiation pattern. Next, with a single RF path on and with constant transmit power and frequency, the receiver-to-transmitter angle was varied from −60° to 60° in 1° increments. The
VGA gain settings were stepped from 3 to 9 dB (i.e., 6 dB of gain tuning range) in 1 dB increments at each angle and the received power was measured. Figure 2.35 shows a single antenna’s radiation pattern in one quadrant under these settings.

![Diagram](attachment:image.png)

**Figure 2.34.** Super-pixel beam steering radiation patterns compared with single antenna radiation pattern, (a) beam steering from −18° to 0°, (b) beam steering from 0° to +18° [66, 67].
Figure 2.35. Single antenna beam pattern for various VGA gain settings [66, 67].

The active imaging test setup (Figure 2.36) using a W-band transmitter to illuminate the receiver chip was similar to the setup of Figure 2.33, with the addition of an object and a mechanical step positioner placed between the signal source and the receiver chip. The illumination frequency was 90 GHz, the output power of the illuminating source was set to a level to prevent saturation of the detector, the integration time was 20 ms, the distance from the illuminating source to the object was 20 cm, and the object-to-receiver distance was 30 cm. The receiver chip was programmed to 0 degree beam steering and VGA gain was set to maximum. The object was then stepped in the x and y directions in 4 mm increments and the detector outputs were recorded 10 times at each position and then averaged. The total scan time was greater than one hour due to the manual operation of the stepper. Without focusing optics, an imaging receiver employing wide beam width on-chip antennas will have a difficult time forming a passive image of a practical sized object, as the wide antenna beam angle will pick up background noise power from adjacent objects, corrupting the noise power of the desired object to be imaged. To overcome this problem, an active test with an illuminating source was used to test the imaging receiver. Figure 2.37 shows the images from the outputs of four individual
super-pixels, obtained by 4 mm spatial sampling, compared with the image obtained by combining the four overlapping super-pixels’ data, resulting in 2-mm sampling. It can be visibly seen that the 2-mm oversampled image results in a sharper spatial resolution.

Figure 2.36. Measurement setup for active imaging [66, 67].
2.5 Summary

This chapter starts with a discussion about the difficulties to achieve a high performance OCA in (Bi)CMOS. Later, a few novel OCA designs showing improved radiation performance are presented. The designs are mainly separated into two categories based on where the ground plane is located. One category is the OCA designs with a ground at the lower metal layer, such
that the insulator layer acts as the antenna substrate. In this way, the major loss is resulting from the ohmic losses of the metal due to the extremely thin antenna substrate. The other category is the OCA designs with the ground plane below silicon. Due to the high permittivity and low resistivity of silicon, most of the energy is dissipated inside the silicon substrate as losses.

It is confirmed that the performance of antenna designs with a ground plane at the lower metal layer is much more robust than those with the ground below silicon, although the latter designs show wider impedance bandwidth and gain bandwidth, in general, when assuming a finite die size. Without a shielding ground plane at the lower metal layer, unstable radiation performance was observed because of the strong coupling between antenna and the high permittivity substrate.

It is noteworthy that for sub-mm-waves and THz OCA design, the performance of the antenna with a ground above silicon will be improved due to the thicker electrically thickness of the silicon dioxide layer. On the contrary, when frequency increases, the OCA with a ground plane below silicon will suffer more from dielectric losses and surface waves inside the silicon. Therefore, it can be inferred that for sub-mm-waves and THz OCA design, a ground at the lower metal layer is strictly demanded.
Chapter 3  Using a Metasurface as a Reflector in On-chip Antennas

Design

3.1 Introduction

High impedance surfaces (HISs) are one kind of metasurfaces based on a grounded dielectric substrate with planar metallic periodic structure. The term HIS was originated in [69], based on the “mushroom” resonant structure on a metallic ground plane. The properties of a HIS can be found by resorting to the concept of equivalent surface impedance, which is defined by the relation between the tangential electric field and the tangential magnetic field. In practice, for HISs made of periodic structures the dyadic form of the surface impedance $Z$ is defined by the relation $\mathbf{E}_t = \mp \mathbf{Z} \cdot \mathbf{n} \times \mathbf{H}_t$, where $\mathbf{E}_t$ is the averaged tangential electric field, $\mathbf{H}_t$ is the averaged magnetic field on the surface and $\mathbf{n}$ is the unit vector perpendicular to the surface. For plane wave incidence, in many practical cases one can deal with a scalar surface impedance $Z_s = E_t / H_t$. The following properties of HIS were reported in [69], including (i) suppression of propagating surface wave along the substrate (ii) in-phase full reflection of incident plane wave, which also be referred as artificial magnetic conductor (AMC). An ideal surface with very high surface impedance ($Z_s \to \infty$) implies that the surface is acting as an open circuit approximately and would fully reflect incident waves for all angles of incidence, and will not support any surface wave. In practice, the equivalent surface impedance of an artificial composite layer is a function of both the frequency and the transverse wave number (associated to temporal and spatial dispersion). Thus, the two properties above do not occur at all frequencies and for all angles of wave incidence. In [69], the HIS is modeled as a parallel resonant circuit system. As
the structure encounter electromagnetic waves, currents are induced in the top metal sheet and charges accumulated on the ends of the plates, which is described as a capacitance. And as the charges flow around the long path around and the vias and bottom plate, an inductance is used to represent the currents associated with the magnetic field. This parallel resonance model exhibits a high impedance at its resonant frequency, which is given by 

$$\omega_0 = 1 / \sqrt{LC}.$$  

In this work, metasurfaces are investigated for OCA design in two different ways, one is using the metasurface as a reflector and the other is to use the metasurface as the radiating antenna directly. This chapter will be focused on using the metasurface as a reflector below the OCA. The idea of using the metasurface made by the dogbone shaped structure, as shown in Figure 3.1, in OCA is studied. According to [70], this type of dogbone shape metasurface shows a couple of design advantages over other type of metasurfaces, (i) it is fully planar and no via is required and (ii) the resonant frequency is easily estimated by an electro- and magnetic static model simply from its geometrical parameters as long as the metasurface substrate is thin enough such that the fringing field effect and the coupling between adjacent dogbone could be neglected.

![Figure 3.1. Dimension of unit cell of dogbone shape element.](image-url)
Placed below a radiating antenna, e.g., a dipole or a folded dipole as in [25, 71-73], AMCs have been proven effective in enhancing the antenna gain [25], widening the input impedance bandwidth and reducing the antenna profile [73]. Recently, artificial magnetic conductor (AMC) or high impedance surface (HIS) types of metasurfaces have been utilized for the OCA designs [32, 74, 75]. Realized in the interlayers of the metal stack-up in silicon, these periodic structures act as reflection planes below the antenna to improve the antenna bandwidth and gain. Different types of AMCs, including the snowflake structure in [32] and the Jerusalem cross structure [75] have been employed in OCA design. In each of these metasurfaces, the ground plane is placed below the lossy silicon, as shown in Figure 3.2(a). In this chapter, besides using the AMC with ground plane below the silicon, investigation was conducted to implement the AMC with the ground plane at the lower metal layer of the CMOS process. It is found that for a metasurface design with a ground plane on the bottom metal layer, i.e., M1, while the periodic structure is placed in the metal interlayers between top and bottom layer as shown in Figure 3.2 (b), the zero-phase crossing will not mostly occur and the metasurface will not exhibit a high impedance property. The fundamental reason of this phenomenon will be discussed. The guideline of how to improve the AMC performance is provided.
Figure 3.2. Two different stack up schemes to use an AMC in CMOS process: (a) the ground plane is below silicon and the wave penetrates the substrate, (b) extremely thin (10 µm) AMC with the ground plane on the lowest metal layer (M1) providing shielding from the silicon substrate.

3.2 Metasurface Reflector with Ground Below Silicon

3.2.1 Simulation of Metasurface Reflector with Ground Below Silicon

In [70], it was shown that the metasurface composed of periodic dogbone elements possesses a magnetic resonance. While it is operating near the magnetic resonance frequency, the metasurface is acting as an artificial magnetic conductor (AMC) whose condition is characterized, assuming low losses, by a vanishing phase of the reflection coefficient for a plane wave with orthogonal incidence. According to the image principle, placing a horizontal source current over a perfect magnetic conductor (PMC) induces an image current having the same direction as the source current on the other side of PMC, which inherently enhance the antenna radiation.

The equivalent circuit model in [70] provides an approximate method to calculate the magnetic resonant frequency of a grounded dogbone. Figure 3.3 (a) illustrates the simulation setup of plane wave incidence over a metasurface made by a ground-backed infinite array of dogbones. Periodic boundary conditions, including two perfect conductor conductor (PEC) surfaces in the y-z plane and two perfect magnetic conductor (PMC) surfaces in the x-z plane are applied in the simulation. It should be noted that the dogbones are placed on M5 while the ground is below the silicon substrate, with the stackup information detailed in Figure 3.2 (a). The phase and magnitude of the plane wave reflection coefficient are shown in Figure 3.3 (b). The dimensions of the dogbones are optimized to make the zero-phase reflection occur at 94 GHz,
which is very close to the metasurface’s magnetic resonance frequency. It can also be seen that at 94 GHz, the magnitude of the reflection coefficient has its minimum value. It is because at the magnetic resonance the current reaches its peak value and thereby induces the maximum ohmic losses. The optimized dimensions of the unit dogbone element are as follows, $A = 0.4 \text{ mm}$, $B = 0.25 \text{ mm}$, $A1=50 \mu \text{m}$, $B1 = 150 \mu \text{m}$, $A2 = 150 \mu \text{m}$, and $B2 = 90 \mu \text{m}$ according to Figure 3.1.

![Simulation setup of plane wave normal incidence over AMC made by dogbone elements with periodic boundary condition.](image)

**Figure 3.3.** (a) Simulation setup of plane wave normal incidence over AMC made by dogbone elements with periodic boundary condition, which comprises two perfect electric conductor surface in +x and –x plane and two perfect magnetic conductor surface in +y and –y plane (b) Amplitude and phase of reflection coefficient of plane wave incidence over AMC made by dogbone.

### 3.2.2 Simulation of Antenna with a Metasurface Reflector and Ground Below Silicon

Figure 3.4 shows the configuration of a dipole antenna above a metasurface made of a ground-backed dogbone array. The dimensions of the dogbones are as that in Figure 3.3 (b). The
The dipole antenna is on the topmost metal layer-M6 while the dogbone layer is on M5. These two layers, M5 and M6, are used because they are thicker than the others, which mean less conduction loss. Below the silicon substrate, it is assumed that there is a ground plane, which in practice could be realized by paint, an interconnection layer in the chip package or in the printed circuit board (PCB). The total area of the proposed antenna is 2 mm × 1.2 mm. The dipole length is 660 µm, which is designed to match the ideal 50 Ω lump port around 90 GHz.

![Diagram of dipole antenna with AMC and ground plane](image)

Figure 3.4. Simulation setup of a dipole antenna above AMC made by dogbone elements while the ground plane is below the silicon substrate.

Figure 3.5 shows the accepted gain of a dipole over the dogbone metasurface illustrated in Figure 3.4 versus frequency. The phase of reflection in Figure 3.3 (b) is instructively plotted together so as to compare with the gain’s dependency over frequency. It can be observed that at the zero-phase reflection frequency (94 GHz), the gain reaches its peak value, +1.2 dB. Considering the losses in the chip environment, the results show that this design provides a good gain. Figure 3.6 (a) shows the accepted gain pattern in the E and H planes of the dipole antenna.
in Figure 3.4 at 94 GHz. Due to the perfect symmetry of the simulated structure, the gain pattern is symmetric in both the E and the H planes. The antenna input reflection coefficient is plotted in Figure 3.6 (b). The –10dB input impedance bandwidth is 12 GHz, from 85 GHz to 97 GHz.

Figure 3.5. Broadside gain ($\theta = 0^\circ$) of the dipole antenna above a grounded dogbone layer and the phase of reflection as in Figure 3.4.
3.3 Metasurface Reflector with Ground at M1

3.3.1 Simulation of a Metasurface Reflector with Ground at M1

In [32, 74, 75], the ground plane of these metasurfaces is placed below the lossy silicon to ensure the typical zero-degree phase crossing property of AMCs [69, 76]. However, when the ground plane of the metasurface is below the silicon substrate, as in Figure 3.2 (a), the OCA performance suffers from the excitation of the surface wave(s) in the silicon substrate, leading to impairments such as losses, coupling, and higher sensitivity of antenna performance on the die’s dimensions. It is worth mentioning that planar electromagnetic bandgap (EBG) structures, such as the mushroom geometry [69], can provide a stop-band for surface waves propagating inside the substrate. However, bandgap structures typically require the periodic structure to be connected to the ground plane with vias, to interact with the vertical electric field of the TM surface waves [69]. In the recent years, the through-silicon-vias (TSVs) have become available.
in customized CMOS technologies, and could meet the demand discussed above. However, in most standard processes, TSVs are still not available. Furthermore, at mm-wave frequencies the silicon substrate is rather thick in terms of the wavelength, hence full stop bands are rather difficult to be realized without additional die back-grinding.

According to the previous discussion in Chapter 2, a ground plane above the silicon is demanded for a robust OCA design. For a metasurface design with a ground plane on the bottom metal layer, i.e., M1, while the periodic structure is placed in the metal interlayers between top and bottom layer as shown in Figure 3.2 (b), the zero-phase crossing will not mostly occur and the metasurface will not exhibit high impedance property. The metasurface designs fail to qualify as AMCs just because of this fundamental limitation arising from the losses in the metasurface when extremely thin substrate thickness is required. For example it is shown in [74] that at the millimeter-wave frequencies, it is hardly possible to design an AMC in Bi(CMOS) chip environment, while using SiO2 as substrate, which has a thickness thinner than $\frac{\lambda_d}{200}$.

Here we show a very thin metasurface made by dogbones at millimeter waves using the stackup of a 180 nm BiCMOS technology in Figure 3.2 (b). The geometry of the dogbone-based metasurface unit cell is shown in Figure 3.1 with the following dimensions (all in $\mu$m): $A = B = 300$, $A1 = 50$, $B1 = 150$, $B2 = 20$, $A2 = 250$, and $h = 9.5$. Conductors are made of copper with $\sigma = 5.8 \times 10^7$ S/m, and have a thickness of 2 $\mu$m for the dogbone and 0.5 $\mu$m for the ground plane). The relative permittivity of the substrate (between the dogbone conductor and the ground plane) is $\varepsilon_r = 4$, typical of the oxide layers in a CMOS technology. From FEM based full wave simulations, the resonance frequency, corresponding to a real reflection coefficient, is $f'_0 \approx 152$ GHz. From the full wave simulation results shown in Figure 3.7, it can be observed
that around the resonant frequency (152 GHz), the phase of the reflection coefficient ($\Delta \Gamma$) does not cross $0^\circ$.

![Graph showing phase of reflection coefficient](image)

**Figure 3.7.** Phase of the reflection coefficient of a plane wave incidence over an extremely thin metasurface made by dogbone elements at mm-wave.

### 3.3.2 Simulation of Antenna above a Metasurface Reflector with Ground at M1

Figure 3.8 shows the simulation setup of a dipole antenna above the metasurface made by dogbone elements as described in *Section 3.3.1*, with the ground at the lower metal layer M1. The dogbone elements are in the same dimension as that in Figure 3.7. It can be observed from the antenna simulated gain shown in Figure 3.9 that the gain is below $-15\text{dBi}$ over the frequency range of interest. It means that since the AMC phenomenon does not occur from the metasurface, the metasurface cannot improve the antenna gain on the top, compared with the case when only the dipole is above the ground plane at M1, without a metasurface in between.
Figure 3.8. Setup of a dipole antenna above a metasurface made by dogbone elements, with the ground plane at the lower metal layer, M1.

Figure 3.9. Broadside gain versus frequency for a dipole antenna with and without a metasurface made by dogbones at below. The ground plane is at M1 as shown in Figure 3.8.

3.3.3 Transmission Line Model of Extremely Thin AMCs

Losing AMC property does not only inhibit realizations at millimeter waves, but also at microwave frequency regime, depending on the thickness of the metasurface. In Figure 3.10,
based on a metasurface made of dogbone-shaped conductors with the same dimension used in [77], over a grounded dielectric substrate, the phase of the reflection coefficient for plane wave normal incidence are plotted versus frequency. The results are obtained for different substrate thickness $h$ (0.1 mm, 1 mm and 2 mm) by full wave simulations using HFSS. For each thickness considered, the results are compared considering two cases: using copper and using perfect electric conductor (PEC) as metal. It is noticed that when the substrate thickness is electrically “thick” ($h = 1 \text{mm and 2 mm}$), losses on the metal do not alter the zero-phase crossing of the AMC reflection coefficient. As a contrast, when the thickness becomes very electrically-thin ($h = 0.1 \text{mm}$), the zero-phase crossing disappears if the AMC is lossy. This clearly implies that the AMC phenomenon is more susceptible to the losses when the metasurface has a thinner substrate compared to a thicker substrate.

Therefore, it is vital to study the effects of losses in very electrically-thin AMC designs and understand certain limiting conditions. In terms of this motivation, in this study we employ an equivalent lumped circuit model of the AMC connected to a transmission line (TL) modeling the plane wave propagation in free space and we derive the necessary criteria to be met to achieve a guaranteed AMC property at the resonance frequency of the metasurface.
Figure 3.10. Phase of the reflection coefficient of a plane wave incidence over the metasurface in [77] for various metasurface thickness when metals are copper ($\sigma = 5.8 \times 10^7$ S/m) or PEC ($\sigma \rightarrow \infty$).

The plane wave reflection at a metasurface is described using the TL and lumped circuit model. Indeed, for very thin metasurfaces, when the thickness ($h$) is much smaller than the in-plane periods, higher order Floquet harmonics, associated to the in-plane periodicity, have impact on the metasurface reflection greatly due to the coupling between the patterned metal surface and the grounded substrate. In other words, higher Floquet modes are able to be reflected by the ground plane and coupled again to the periodic patterned conductor, due to the very thin thickness. Hence, the method modeling the periodic structure with RLC elements while using standard TL model for the substrates and using a shunt lumped impedance for the frequency selective surface (FSS) on top, as those in [73, 78], may lose validity since they are based on a single dominant TL mode propagating through the substrate and its reflection from the bottom ground plane. In those models, the effect of higher order Floquet waves reflected from the ground plane may be taken into account by a further modification of the reactance of the lumped
shunt load characterizing the FSS. This further modification is in general implicit since the impedance of the shunt load is found by curve fitting.

This is the reason why here we model the metasurface with a simple network of lumped elements (as in Figure 3.11), also justified by the assumption of very thin substrate thickness, which makes use of the lowest resonances of the reflection coefficient. Indeed, here we are mainly interested in extremely subwavelength thin metasurfaces, and therefore higher order Floquet waves still play a role. Their effect is included in the lumped circuit, which is here based on a low order expansion of the TL effective load [53], found by curve fitting.

Accordingly, the circuit model of the thin metasurface comprises a lumped network, loading the transmission line as depicted in Figure 3.11, and it is composed of a capacitor $C$ in parallel to an inductor $L$ and a series resistor $R$. As mentioned previously, here we assume that the losses are mainly due to the ohmic losses in the conductors, considering the substrate is extremely thin. It is observed by full wave simulations that the losses on AMC are significantly

Figure 3.11. (a) Plane wave incidence over a high impedance surface, (b) the equivalent model for a plane wave incident on the metasurface represented as a lumped resonant load. The resistance models losses.
dominated by the ohmic losses when the substrate is extremely thin. Therefore in our model, the
loss resistance $R$ is naturally on the inductor branch of the parallel LC resonator. A similar
parallel RLC lumped model was proposed in [79] to model the loss property by checking the
reflection coefficient of plane wave incidence over a periodic array of the reflectarray elements.

According to the lumped model in Figure 3.11, the impedance of the metasurface is

$$Z_{MS} = \frac{1}{j\omega C} || (j\omega L + R) = \frac{j\omega L + R}{1 - \frac{\omega^2}{\omega_0^2} + j\omega RC}, \quad (3.1)$$

where $\omega_0 = 1/\sqrt{LC}$, which is the undamped natural angular frequency of the $RLC$ resonator.

The quality factor of the $RLC$ resonator at $\omega_0$ is

$$Q = \frac{\omega_0 L}{R} = \frac{1}{\omega_0 RC} = \sqrt{\frac{L}{C}} \frac{1}{R} = \eta \frac{1}{R}. \quad (3.2)$$

This implies that at the undamped natural resonant frequency, the lossy metasurface shows
capacitive, instead of inductive impedance. In the presence of losses, the actual resonance
frequency ($\omega'_0$) is found by imposing $\text{Im}(Z_{MS}) = 0$, which occurs at a slightly shifted and
smaller frequency

$$\omega'_0 = \omega_0 \sqrt{1 - \frac{1}{Q^2}}. \quad (3.3)$$

At this frequency the metasurface shows a purely real impedance

$$Z_{MS}(\omega = \omega'_0) = \eta R = \frac{L}{C} \frac{1}{R} = \frac{\eta^2}{R}. \quad (3.4)$$
which could be larger or smaller than $Z_0$. Note that (i) this tells us that when loss effect is taken into account, the resonant frequency is smaller than that of the lossless case, (ii) if $Q < 1$, there is no real frequency $\omega_0'$ to satisfy $\text{Im}(Z_{MS}) = 0$. This hereby implies that $Q$ determines whether the AMC phenomenon occurs or not in the metasurface.

3.3.4 Threshold Condition of AMC Occurrence

The reflection coefficient of a metasurface is usually considered just above the top surface of metasurface. According to the model in Figure 3.11, it is

$$\Gamma = \frac{Z_{MS}(\omega) - Z_0}{Z_{MS}(\omega) + Z_0}$$

(3.5)

Here, $Z_0$ is the wave impedance of the medium above the metasurface (e.g., free space), which also accounts for the dependence on polarization and incidence angle of the plane wave. And when evaluated at the exact resonance frequency we have

$$\Gamma(\omega_0') = \frac{\eta^2}{R} - \frac{Z_0}{2} \frac{1 - \frac{Z_0}{RQ^2}}{1 + \frac{Z_0}{RQ^2}} = \frac{1 - \frac{Z_0}{\eta Q}}{1 + \frac{Z_0}{\eta Q}}$$

(3.6)

which is purely real. When $Z_{MS}(\omega_0') = RQ^2 = \eta^2 / R$ is larger than $Z_0$, $\Gamma(\omega_0')$ is a positive real number, i.e., the reflection phase $\angle \Gamma(\omega_0') = 0^\circ$, similar to the reflection phase from an AMC surface [69]. On the other hand, when $Z_{MS}(\omega_0')$ is smaller than $Z_0$, $\Gamma(\omega_0')$ is a negative real number ($\angle \Gamma(\omega_0') = 180^\circ$), therefore the reflection phase is similar to that of an electric conductor.
surface. If one has \( Z_{MS}(\omega'_0) = Z_0 \) than one may even have \( \Gamma(\omega'_0) = 0 \), however it should be noted that this condition corresponds to a perfectly matched layer leading to no reflection.

To have a phase \( \angle \Gamma(\omega'_0) = 0^\circ \), \( \Gamma(\omega'_0) \) has to be a positive real number, and thus the resistance has to be such that

\[
R < \frac{L}{CZ_0} = \frac{\eta^2}{Z_0}
\]  

or in terms of the quality factor, \( Q > \frac{Z_0}{\eta} = Z_0 C \omega_0 \), which can be also written as

\[
Z_0 < \frac{\eta^2}{R}
\]  

where the left side is a constant value for free space at a fixed incidence angle and polarization, and the right hand side depends only on the AMC parameters. Equation (3.8) indicates that one needs to increase the ratio \( \eta \) between the \( L \) and \( C \) and/or decrease \( R \). Equation (3.7) or (3.8) defines a threshold condition to estimate the existence of the AMC phenomenon from a general metasurface.

As shown in Figure 3.12, the total system could be separated into two parts, one is the series \( R \) and \( L \) and the other one consists of a parallel circuit between \( C \) and \( Z_0 \). The quality factor of those two parts at the undamped natural resonant angular frequency (\( \omega_0 \)) are \( Q_L = \omega_0 L / R \) (corresponds to the previously introduced \( Q \) in (3.2)) and \( Q_C = Z_0 C \omega_0 \). Using \( Q_L \) and \( Q_C \), the threshold condition (3.7), or (3.8), could be equivalently expressed as

\[
Q_L > Q_C
\]  

(3.9)
which reveals that for a metasurface to act as an AMC, the quality factor of the series $RL$ has to be larger than the quality factor for the parallel $C, Z_0$.

![RLC Circuit](image)

Figure 3.12. The resonating circuit where RLC circuit modeling the metasurface is loaded by the wave impedance $Z_0$.

From (3.6), one could also observe that the magnitude of the reflection coefficient is smaller than unity because of losses. When the AMC condition (3.7) is satisfied, an equivalent larger $R$, larger $C$ and smaller $L$, i.e., smaller $Q\eta$, will lead to a lower dip in the magnitude of reflection coefficient curve, i.e., more power is dissipated as losses in the metasurface. This seems to follow the intuitive concept that $R$ is directly related to the losses on the metasurface, and it coherently indicates that these losses would affect the magnitude of the reflection coefficient around the resonance. However, more precisely, once AMC property occurs, the exact condition to minimize absorption is to surpass the condition

$$\eta Q \gg Z_0$$

(3.10)
as much as possible. Indeed, total absorption occurs at the threshold condition $\eta Q = Z_0$, corresponding to $\Gamma(0) = 0$. Also it means that an ideal perfect magnetic conductor (PMC), i.e., $\Gamma(0) = 1$, would never occur as long as there are losses in the metasurface.

When the AMC is placing below an antenna as a reflector, the magnitude of the reflection coefficient in the resonant frequency region determines how much energy the AMC reflects back to the antenna on the top, instead of dissipated as the losses in AMC. If the magnitude of reflection is too small at the frequency region close the resonance, even although the zero-phase crossing appears in the phase, the effect of AMC will be partially weakened. Thereby, it is useful to define a minimal acceptable reflection magnitude at the resonance frequency. If one consider a $-3$dB magnitude of reflection coefficient as the minimum acceptable level, according to (3.6), the condition

$$3Z_0 < \frac{L}{CR}$$

(3.11)

must be satisfied. (3.11) could be also considered as a threshold condition to achieve a highly effective AMC.

### 3.3.5 Illustrative AMC Cases

Here we analyze three different representative cases at both microwaves and millimeter waves. The examples stress the importance of the threshold condition and the analysis in the previous section, showing also cases where the AMC property does not occur.

First of all, we consider an AMC made of a periodic array of dogbone shaped structure shown in Figure 3.1, with dimensions as follows (all in mm), $A = 16$, $B = 11.7$, $BI = 4$, $AI = B2$
= 0.8, \( A_2 = 14 \), and \( h = 2 \). The metal is copper \((\sigma = 5.8 \times 10^7 \text{ s/m})\) and the substrate is foam with a relative dielectric constant \( \varepsilon_r \approx 1 \), and considered as lossless. Using the finite element method (FEM) full wave solver HFSS, we found the resonance frequency \( f'_0 \approx 6.5 \text{ GHz} \), by observing a real-valued reflection coefficient (above the top metal layer), in a simulation with periodic boundary conditions and a plane wave incidence. Note that in this case \( h \approx \lambda_d / 20 \) at \( f'_0 \).

The circuit network parameters of \( R \), \( C \) and \( L \) are extracted by curve-fitting the full wave simulation results of the reflection coefficient of plane wave incidence over the top surface of the metasurface using the FEM-based full wave simulator. The circuit simulator Agilent ADS is used to search for the optimal \( RLC \) values to best fitting the curve from full wave simulation. The fitting was carried out considering the frequency range close to zero-phase crossing of the reflection phase and the reflection minimum \((\min|\Gamma(\omega)|)\). The RLC values obtained are \( R = 25.6 \text{ m}\Omega, L = 0.739 \text{ nH}, C = 0.82 \text{ pF} \). From (3.2), the quality factor is fairly large: \( Q = 1115 \) (thus \( f'_0 \approx f_0 \)).

As a comparison, in Figure 3.13, the phase and magnitude of reflection coefficient from full wave simulation and from the extracted TL model are plotted together, as a confirmation that these values provide the correct metasurface response over a large frequency range. It can be seen that for this case, the zero-phase crossing of the reflection coefficient exists at the resonance frequency. Meanwhile, according to the RLC values extracted above, one has \( \left( \eta^2 / Z_0 \right) = 2.79 \Omega \) and \( R = 0.03 \Omega \) and therefore the threshold condition is clearly satisfied since \( R \ll \eta^2 / Z_0 \).
Figure 3.13. Comparison of the reflection coefficient from the full wave simulation and the synthesized method using the equivalent model of the metasurface made of dogbones for the case when losses are moderate.

It should be noted that the results of reflection coefficient obtained by full wave simulation is evaluated at the top-most surface of the metasurface, i.e. the equivalent effective location of the metasureface is chosen at its top. In reality the metasureface is not infinitely thin, therefore the equivalent ‘effective’ vertical location of the metasurface, i.e. the reference place for the reflection coefficient, should be somewhere inside the metasurface, more specifically between the ground plane and the plane made by the periodic structures.

Similarly, for a metasurface made of dogbone shaped conductors on the a substrate with different thicknesses as shown in Figure 3.10, the values for the lumped element are retrieved
using the same method discussed above respectively and shown in Table 3.1. It can be observed that the values of $R$, $L$ and $C$ vary when the metasurface thickness decreases. Indeed, according to the electrostatic and magnetostatic model as discussed in [70] for dogbone shape element, the total capacitance of the dogbone is associated with the charges accumulated on its lateral arms and the equivalent capacitor could be estimated by the parallel plate capacitor model for the two lateral arms of dogbone. Meanwhile, the inductance of the dogbone is dominated by the central section of the dogbone. According to these two models, the capacitor could become larger while the inductance would be smaller when the thickness of a metasurface decreases. This actually matches with the trends as shown in Table 3.1.

In terms of the right hand side of threshold condition in (3.8), both the numerator and the denominator decrease when the metasurface becomes thinner. Meanwhile, the numerator decrease much faster than the denominator. Considering the left hand side of the threshold condition (3.8) is a fixed value for free space under normal incidence, this observation provides the insight that a thinner thickness $h$ would make the metasurface more susceptible to lose the AMC property.

Table 3.1. Values of extracted RLC lumped elements for dogbone normal incidence case, with different thickness as shown in Figure 3.10.

<table>
<thead>
<tr>
<th>Thickness $h$ (mm)</th>
<th>$R$ (mΩ)</th>
<th>$L$ (pH)</th>
<th>$C$ (pF)</th>
<th>Satisfy Threshold (3.8)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1</td>
<td>3.94</td>
<td>11.33</td>
<td>39.8</td>
<td>No</td>
</tr>
<tr>
<td>1</td>
<td>20.35</td>
<td>306.8</td>
<td>1.6</td>
<td>Yes</td>
</tr>
<tr>
<td>2</td>
<td>26.3</td>
<td>699</td>
<td>0.8</td>
<td>Yes</td>
</tr>
</tbody>
</table>
Additionally, we consider the case of very thin metasurface at millimeter waves as described in Section 3.31. For this case, the metasurface loses the AMC’s zero-phase crossing behaviour because of ohmic losses. Indeed, considering the current flowing on the central part of the dogbone, we recognize that this is equivalent to the case of a microstrip line over a very thin substrate. And it is known that the ohmic losses may be overall significant for very thin substrate thickness [80]. From FEM based full wave simulations, the resonance frequency, corresponding to a real reflection coefficient, is \( f_0' \approx 152 \text{ GHz} \). In this case the substrate thickness is \( h \approx \lambda_d / 104 \) at 152 GHz, which is very subwavelength and turns out to be a key factor to affect the AMC performance. The extracted circuit model parameters values are \( R = 21.9 \text{ m\Omega}, L = 0.942 \text{ pH}, C = 1.16 \text{ pF} \), which lead to a quality factor \( Q = 40 \) (thus \( f_0' \approx f_0 \)).

From the full wave simulation results shown in Figure 3.7, it can be observed that around the resonant frequency (152 GHz), the phase of the reflection coefficient \( \angle \Gamma \) does not cross 0°. In contrast to the previous case at microwave frequencies, the circuit parameters of this case do not satisfy the condition in (3.8) since one has \( \left( \eta^2 / Z_0 \right) = 0.0022 \Omega \) which is much smaller than \( R = 0.022 \Omega \).
Figure 3.14. Comparison of the reflection coefficient from the full wave simulation and the synthesized method using the equivalent TL model of the metasurface made of dogbones for the case when there are *strong* losses.

Besides, we consider a very thin metasurface at millimeter waves, however of a “mushroom” type [69, 72] made of square patches with vias, implemented in BiCMOS technology as in Case B. The geometry of the single unit cell is shown in Figure 3.15 with the following dimensions (all in µm): \( A = 455 \) and \( B = 450 \). The via between the metal patch and the ground plane has a square cross section with size of \( 30 \, \mu \text{m} \times 30 \, \mu \text{m} \). The substrate thickness is \( h = 9.5 \, \mu \text{m} \), as in the dogbone cases at mm-waves.

Figure 3.15. Dimension of unit cell of mushroom shape element.
From full wave simulations, the resonance frequency, corresponding to a real reflection coefficient, is \( f_0' \approx 152 \text{ GHz} \). In this case the substrate thickness is \( h = \frac{\lambda_d}{104} \) at 152 GHz, which is also very subwavelength, similar to the dogbone cases at mm-waves. The extracted circuit model parameters values are \( R = 166.3 \text{ m}\Omega \), \( L = 6.98 \text{ pH} \), and \( C = 155.22 \text{ fF} \), which lead to a quality factor \( Q = 40.3 \). Similar to the dogbone cases at mm-waves, the AMC property is not observed. In Figure 3.16, there is a good agreement between the synthesized RLC and full wave simulation results of the reflection coefficient shown. Therefore the RLC model's applicability to model other types of metasurfaces is confirmed.
3.3.6 AMC Under Oblique Incidence Wave

Since both the surface impedance of a metasurface and the wave impedance of an incoming plane wave, have strong dependence on the angle of incidence and the polarization of the incident wave, the response of the metasurface, in particular if the AMC property occurs, should be particularly studied for oblique incidence cases.

In Table 3.2 and Table 3.3, we show the values of the extracted circuit parameters for the dogbone-based metasurface at the microwave frequencies according to different incidence angels for TE- and TM-polarized plane wave respectively. We also indicate the quality factor of the total system ($Q_{Tot}$) comprising the metasurface and the free space medium on the top at the resonant frequency ($f_0$) in the tables, which is evaluated as

$$Q_{Tot} = \frac{Q_C Q_L}{Q_C + Q_L}.$$  \hspace{1cm} (3.12)

The free space wave impedance $Z_0$ is expressed by $Z_0 = \sqrt{\mu_0 / \epsilon_0} / \cos \theta$ for TE-polarized and $Z_0 = \sqrt{\mu_0 / \epsilon_0} \cdot \cos \theta$ for TM-polarized plane wave, where $\theta$ is the angle of plane wave incidence. The quality factor of the $RL$ branch is evaluated as $Q_L = \omega_0 L / R$ while the quality factor of the parallel $Z_0, C$ branch is evaluated as $Q_C = \omega_0 CZ_0$.

Figure 3.17 shows the full wave simulation results for the phase of the reflection coefficient of TM-polarized plane wave incidence over the dogbone-based metasurface for...
different incidence angles. The dimension of dogbone follows that in the microwave dogbone case. We can see that the AMC property occurs at normal angle of incidence and larger angles, but it disappears when the incidence angle is 80 degrees, for a TM polarized wave. The same trend is not observed in the case of TE-polarization (Table 3.3). For this case, the AMC property is observed for any incidence angles.

Applying the lumped model discussed above, the value of $R$, $L$ and $C$ is retrieved for different incidence angles, for both TM and TE-polarized incident waves, as shown in Table 3.2 and Table 3.3. Curve fitting between the TL model and full wave simulations yields to a very good matching of the reflection coefficient over a wide frequency range, for every incident angle and polarization, further demonstrating the effectiveness of the proposed lumped model to represent metasurfaces even at oblique incidence. In particular, the values of $R$, $L$, $C$ show angular dependency for TM polarization but not for TE polarization. Similar angular dependency of the lump elements’ value was also observed in modeling frequency selective surface using

![Figure 3.17. Phase of the reflection coefficient of a plane wave normal incidence over the metasurface made by dogbones in the microwave case of dogbone for various oblique incidence angles.](image-url)
lumped model [81-83]. For all the angles considered, the quality factors \((Q_L, Q_C\) and \(Q_{Tot}\)) are also shown in the tables. For the TM-polarized case, the threshold condition (3.8) is satisfied for incidence angles equal or less than 75 degrees, and therefore the AMC property occurs, as observed in Figure 3.17. Table 3.2 shows that when the incidence angle is equal to 80 degrees, the threshold condition is not satisfied and correspondingly the AMC property disappears, as also confirmed in Figure 3.17. It is very interesting to observe that the approaching larger angle of incidence the resistance \(R\) decreases, nevertheless the threshold condition (3.8) fails to be satisfied at 80 degrees, proving one more time that it is the ratio \(L/(RC)\) to conclude the appearance of AMC property. Note also that the free space TM wave impedance decreases for larger incidence angles, nevertheless (3.8) fails at larger angles, because the \(L/(RC) = Q_L\eta\) decreases faster with a increasing angle of incidence.

For TE incidence, the threshold condition (3.8) is always satisfied, and indeed the AMC property occurs at any incidence angles. This is a consequence that the values of the equivalent \(R\), \(L\), and \(C\) component, as well as the resonant frequency, are nearly invariant with respect to the incidence angles. In particular, because of its effectiveness, the mode could be used to improve the performance of an AMC metasurface.

It is known that a horizontal antenna radiates a large angular spectrum of TE and TM-polarized plane waves. For a selection of plane wave incidence angles, we show that the proposed lumped model could fit well with the full wave simulation. This further demonstrates the general applicability of the proposed lumped model.
Table 3.2. Oblique incidence case of the dogbone metasurface, TM polarization.

<table>
<thead>
<tr>
<th>Incidence angle (°)</th>
<th>$R$ (mΩ)</th>
<th>$L$ (PH)</th>
<th>$C$ (pF)</th>
<th>$f_0$ (GHz)</th>
<th>$f_0'$ (GHz)</th>
<th>$Q_c$</th>
<th>$Q_L$</th>
<th>$Q_{Tot}$</th>
<th>AMC Threshold Satisfied</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>25.6</td>
<td>739</td>
<td>0.82</td>
<td>6.47</td>
<td>6.46</td>
<td>12.5</td>
<td>1173</td>
<td>12.4</td>
<td>Yes</td>
</tr>
<tr>
<td>15</td>
<td>25.9</td>
<td>665</td>
<td>0.889</td>
<td>6.55</td>
<td>6.55</td>
<td>13.3</td>
<td>1056</td>
<td>13.1</td>
<td>Yes</td>
</tr>
<tr>
<td>30</td>
<td>20.44</td>
<td>460</td>
<td>1.197</td>
<td>6.78</td>
<td>6.78</td>
<td>16.7</td>
<td>959</td>
<td>16.4</td>
<td>Yes</td>
</tr>
<tr>
<td>45</td>
<td>11.57</td>
<td>218</td>
<td>2.296</td>
<td>7.11</td>
<td>7.11</td>
<td>27.4</td>
<td>842</td>
<td>26.5</td>
<td>Yes</td>
</tr>
<tr>
<td>60</td>
<td>3.2</td>
<td>55.2</td>
<td>8.37</td>
<td>7.4</td>
<td>7.41</td>
<td>73</td>
<td>802</td>
<td>67</td>
<td>Yes</td>
</tr>
<tr>
<td>75</td>
<td>0.275</td>
<td>46.2</td>
<td>96.14</td>
<td>7.552</td>
<td>7.552</td>
<td>445</td>
<td>797</td>
<td>285</td>
<td>Yes</td>
</tr>
<tr>
<td>80</td>
<td>0.0696</td>
<td>1.0546</td>
<td>417.4</td>
<td>7.583</td>
<td>NA</td>
<td>1302</td>
<td>721</td>
<td>464</td>
<td>No</td>
</tr>
<tr>
<td>Incidence angle (°)</td>
<td>R (mΩ)</td>
<td>L (pH)</td>
<td>C (pf)</td>
<td>$f_0$ (GHz)</td>
<td>$f_0'$ (GHz)</td>
<td>Qc</td>
<td>QL</td>
<td>QTot</td>
<td>AMC Threshold Satisfied</td>
</tr>
<tr>
<td>---------------------</td>
<td>--------</td>
<td>--------</td>
<td>--------</td>
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<td>----</td>
<td>------</td>
<td>-------------------------</td>
</tr>
<tr>
<td>0</td>
<td>25.6</td>
<td>739</td>
<td>0.82</td>
<td>6.47</td>
<td>6.5</td>
<td>12.5</td>
<td>1173</td>
<td>12.4</td>
<td>Yes</td>
</tr>
<tr>
<td>15</td>
<td>26.28</td>
<td>741</td>
<td>0.819</td>
<td>6.47</td>
<td>6.5</td>
<td>13</td>
<td>1145</td>
<td>12.8</td>
<td>Yes</td>
</tr>
<tr>
<td>30</td>
<td>26.85</td>
<td>738.7</td>
<td>0.82</td>
<td>6.47</td>
<td>6.5</td>
<td>14.5</td>
<td>1118</td>
<td>14.3</td>
<td>Yes</td>
</tr>
<tr>
<td>45</td>
<td>26.285</td>
<td>737.7</td>
<td>0.819</td>
<td>6.48</td>
<td>6.48</td>
<td>17.8</td>
<td>1141</td>
<td>17.5</td>
<td>Yes</td>
</tr>
<tr>
<td>60</td>
<td>26.6</td>
<td>745</td>
<td>0.81</td>
<td>6.48</td>
<td>6.49</td>
<td>24.86</td>
<td>1140</td>
<td>24.3</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Table 3.3. Oblique incidence case of the dogbone metasurface, TE polarization.
3.3.7 AMC Bandwidth

Due to the rapid variation of phase, especially when the metasurface has an electrically-thin substrate, the AMC has a narrow frequency bandwidth. However, for certain applications, a control of the operational bandwidth is desired. To understand how to improve the AMC bandwidth, it is useful to resort to the lumped model in Figure 3.12.

It has been observed that when the zero-phase crossing exists in the reflection coefficient, one could define the AMC bandwidth as in [84], which is the frequency range where the phase of the reflection coefficient is between +90° and −90°. Using the RLC equivalent model, this bandwidth is calculated analytically as

$$BW = \omega_0 \sqrt{\left(\frac{1}{Q_c^2} - \frac{1}{Q_L^2} + 2\right) - 2 \sqrt{1 - \frac{1}{Q_c^2 Q_L^2}}}.$$  \hspace{1cm} (3.13)

where, as before, $Q_c = Z_0 C_0 \omega_0$ and $Q_L = \omega_0 L / R$. The steps of the derivation are provided in Appendix A.

According to the threshold condition (3.9), when $Q_L > Q_C$, the metasurface should show the AMC property. And under the condition $Q_L > Q_C > 1$, one can obtain a real frequency bandwidth, provided by (3.13). When $Q_L < Q_C$, according to the condition (3.9) there is no zero-phase crossing in reflection coefficient, the definition of AMC bandwidth as the frequency range between +90° and −90° does not hold any more and therefore (3.13) is invalidated.

In general, since the resistance in the RLC model is usually much smaller than the free space wave impedance ($R \ll Z_0$) and therefore $\frac{1}{Q_c Q_L} = \frac{R}{Z_0} \ll 1$, one could have an simpler formula for AMC bandwidth as
\[ BW \approx \omega_0 \sqrt{\frac{1}{Q_c^2} - \frac{1}{Q_L^2}}. \]  

(3.14)

From the formula above, we could clearly see that a lower \( Q_L \) of the series \( LR \) due to larger \( R \) could induce a narrow bandwidth of AMC. Assuming the same amount of current flowing on the metasurface a larger \( R \) means more losses. This thereby implies that more loss inside the metasurface would narrow its AMC bandwidth.

If the loss inside AMC is very low, under the assumption that \( Q_L >> 1 \), the above equation could even be simplified as

\[ BW = \frac{\omega_0}{Q_c} = \frac{1}{CZ_0}. \]  

(3.15)

which coincides with the formula of AMC bandwidth reported in [84].

The formulas (3.13) ~ (3.15) are all based on the proposed equivalent lumped model of the metasurface, more specifically, the circuit model matched to the full wave simulations. However, it is necessary to assess how the quality factors determining the metasurface bandwidth evolve with increasing losses in the conductors. Here we start from the case of dogbone structure at microwaves and vary the conductivity of the metal used for the dogbone elements, from that of copper (5.8×10⁷ S/m) to a smaller and smaller value until the AMC property disappear. By varying only the conductivity and keeping the physical dimension of the structure, the value of \( L \) and \( C \) is expected to stay almost constant, because the field profile around the ground plane and the dogbone can be assumed to change negligibly. Meanwhile significant variation should occur in the value of \( R \) if conductivity changes since the resistance \( R \) is mainly due to the ohmic losses on the dogbone and the ground plane below, which is inverse.
proportional to the metal conductivity. Table 3.4 shows the value of $R$, $L$, and $C$ from the curve fitting with the full wave simulation and the resonant frequency ($f'_0$) and the AMC bandwidth observed directly from the phase of reflection coefficient curve.

Table 3.4. Values of the extracted lumped element for the dogbone type of AMC under normal incidence case, with varying conductivity of the metal.

<table>
<thead>
<tr>
<th>$\sigma$ (s/m)</th>
<th>$R$ ($\Omega$)</th>
<th>$L$ (PH)</th>
<th>$C$ (fF)</th>
<th>$f'_0$ (GHz)</th>
<th>AMC BW (GHz)</th>
<th>Threshold</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.8×10^7</td>
<td>0.0297</td>
<td>789.5</td>
<td>742.7</td>
<td>6.57</td>
<td>0.56</td>
<td>Yes</td>
</tr>
<tr>
<td>1×10^7</td>
<td>0.0718</td>
<td>793</td>
<td>740</td>
<td>6.57</td>
<td>0.56</td>
<td>Yes</td>
</tr>
<tr>
<td>1×10^6</td>
<td>0.209</td>
<td>791.2</td>
<td>742.2</td>
<td>6.55</td>
<td>0.56</td>
<td>Yes</td>
</tr>
<tr>
<td>1×10^5</td>
<td>0.933</td>
<td>775.6</td>
<td>760.9</td>
<td>6.52</td>
<td>0.52</td>
<td>Yes</td>
</tr>
<tr>
<td>3×10^4</td>
<td>2.26</td>
<td>766.3</td>
<td>781</td>
<td>6.48</td>
<td>0.26</td>
<td>Yes</td>
</tr>
<tr>
<td>2.7×10^4</td>
<td>2.467</td>
<td>767.9</td>
<td>780.5</td>
<td>6.477</td>
<td>0.2</td>
<td>Yes</td>
</tr>
<tr>
<td>2.53×10^4</td>
<td>2.599</td>
<td>767.77</td>
<td>781.32</td>
<td>6.476</td>
<td>0.052</td>
<td>Yes</td>
</tr>
<tr>
<td>2.52×10^4</td>
<td>2.603</td>
<td>766.77</td>
<td>782.37</td>
<td>NA</td>
<td>NA</td>
<td>No</td>
</tr>
</tbody>
</table>

It can be observed in Table 3.4 that (i) the values of $L$ and $C$ show a small variation while the main variation occurs on the $R$, which agrees with our expectation. Meanwhile the resonant frequency ($f'_0$) also changes due to the change in quality factors. (ii) The AMC property disappears when the conductivity of metal is decreased to 2.52×10^4 S/m. For that case, the left side of the threshold condition (3.8), which is equal to 2.60, becomes slightly greater than the right side (2.5996). This further verifies that the proposed threshold condition could estimate how losses could affect the appearance of AMC property for very thin metasurfaces.

In Figure 3.18, we show the relative bandwidth of the AMC (defined as the ratio between the AMC bandwidth and the zero-phase crossing frequency $f'_0$) versus the extracted $R$ from Table 3.4 in blue square markers. Besides that, we plot the relative bandwidth versus the
resistance using formula (3.13) for two cases when $\sigma = 5.8 \times 10^7$ S/m (red dashed line) and $\sigma = 2.52 \times 10^4$ S/m (black solid line) using the value of $R$, $L$ and $C$ in Table 3.4. One can see that the black solid and the red dashed curves are very close to each other indicating that the modification of the conductivity has slight impact on $L$ and $C$. Moreover, the blue square markers, denoting the designs in Table 3.4, follow the trend determined by the curves obtained from (3.13), i.e, the analytic relation between the bandwidth and the circuit parameters.

![Figure 3.18. AMC relative bandwidth versus $R$: red dashed line -- BW formula (3.13) using $L$, $C$ extracted from the case when metal conductivity $\sigma = 5.8 \times 10^7$ S/m, black solid line -- BW formula (3.13) using $L$, $C$ extracted from the case when metal conductivity $\sigma = 2.53 \times 10^4$ S/m, blue square -- curve fitting from full wave simulation using metal of different conductivity.]

### 3.4 Comparison of Two Extremely Thin On-chip Metasurfaces

In this section, we compare the performance of two potential metasurfaces when using them as on-chip AMCs at millimeter waves, assuming a ground plane at the lowest metal layer (i.e., M1). The cells we studied are dogbones arrayed in a rectangular lattice and gangbusters [85]
arrayed in a triangular lattice. Based on the same CMOS stackup as shown in Figure 2.1, we have observed that the gangbuster structure demonstrates a better performance in terms of a lower frequency to achieve AMC property and a wider AMC bandwidth at the frequency of interest.

The simulated phase of the reflection coefficient ($S_{11}$) for a normal plane wave incidence over the metasurface made up of dogbones are shown in Figure 3.19. When operating at different frequencies, the dogbone dimension parameters are varied and the values are shown in Table 3.5. It can be observed that for the design at 140 GHz, the zero-phase crossing does not occur. When frequency increases, the phase of $S_{11}$ decreases but this trend is reversed at 135 GHz where the phase starts to grow. At the resonant frequency (~ 140 GHz), the reflection coefficient shows a phase of 180 degrees, which means the metasurface acts as an electric conductor rather than a magnetic conductor, since the latter requires a phase close to zero. Meanwhile, the zero-phase crossing is observed for the cases designed to work at 220 GHz and 300 GHz.

One of our objectives is to identify the lowest operating frequency to achieve the AMC property, assuming a extremely thin substrate thickness. A parametric study for each dimensional parameters was conducted (for brevity, results are not shown). To minimize ohmic losses in the dogbone, the width of the center bar of the dogbone ($B_2$) is increased. And the dogbones are placed close together to increase the capacitive coupling between contiguous elements in order to lower the resonant frequency of the structure according to [70]. It is found the lowest frequency to achieve AMC property is 190 GHz. It could be noted that this lowest frequency could be varied in terms of the stackup based on different CMOS processes.
Figure 3.19. (a) Top view for dogbone shape element, (b) the reflection phase of normal plane wave incidence over the metasurface made of dogbones at 140, 220, 300 GHz and 190 GHz (the lowest frequency the zero-phase crossing could be observed).

Table 3.5. Dimensions of dogbone operating at different frequencies (Unit: all in µm).

<table>
<thead>
<tr>
<th>Freq. (GHz)</th>
<th>A</th>
<th>B</th>
<th>A₁</th>
<th>B₁</th>
<th>A₂</th>
<th>B₂</th>
</tr>
</thead>
<tbody>
<tr>
<td>140</td>
<td>346</td>
<td>350</td>
<td>70</td>
<td>340</td>
<td>340</td>
<td>220</td>
</tr>
<tr>
<td>190</td>
<td>276</td>
<td>280</td>
<td>70</td>
<td>270</td>
<td>270</td>
<td>220</td>
</tr>
<tr>
<td>220</td>
<td>226</td>
<td>230</td>
<td>50</td>
<td>220</td>
<td>220</td>
<td>150</td>
</tr>
<tr>
<td>300</td>
<td>166</td>
<td>170</td>
<td>50</td>
<td>160</td>
<td>160</td>
<td>120</td>
</tr>
</tbody>
</table>

Similarly, the reflection phase property is investigated for a metasurface composed by the gangbuster in a triangular lattice, with the top view of the structure shown in Figure 3.20 (a). The reflection phase is shown in Figure 3.20 (b) while the values of each parameter are shown in Table 3.6.

It can be observed in Figure 3.20 that the zero-phase crossing cannot be observed for a design at 140 GHz also for the gangbuster and meanwhile it is observed for the cases working at 220 and 300 GHz. The lowest frequency at which we have observed the zero-phase crossing is 165 GHz, which is 25 GHz lower than that of dogbone. The reason is due to the fact that the
triangular lattice arrangement makes the gangbusters be packed in a higher density compared with the dogbone in rectangular lattice. Therefore more capacitive coupling is established between adjacent gangbusters elements and thereby a lower inductance can be used while keeping the resonant frequency unchanged. This implies also a lower resistance. Establishing a surface with large capacitance seems to be an advantage that the gangbuster metasurface has over the dogbone one.

![Diagram](image)

Figure 3.20. (a) Top view for gangbuster elements in a triangular lattice over an extremely thin grounded substrate, (b) the reflection phase of a plane wave normal incidence over the metasurface made of gangbusters at 140, 220, 300 GHz and 165 GHz (the lowest frequency the zero-phase phase crossing could be observed).

Typically the AMC bandwidth is defined as the frequency range where the phase of the reflection coefficient is between $+90^\circ$ and $-90^\circ$. Here we compare the bandwidth of AMC made of dogbones and gangbusters at 220 and 300 GHz. Figure 3.21 shows the phase of the reflection coefficient for two structures at two frequencies.

For the same structure, the relative bandwidth (defined as the ratio between the AMC bandwidth and the zero-phase crossing frequency) at 300 GHz is larger than that at 220 GHz. This implies that the relative AMC bandwidth will be increased with a higher operating
frequency. Importantly, the bandwidth achieved by AMC composed by gangbusters doubles that by AMC made by dogbones at both 220 and 300 GHz. That means that among the two structures studied here, the gangbuster shows a better performance, at least for the normal incidence case.

Table 3.6. Dimensions of gangbuster operating at different frequencies (Unit: all in µm).

<table>
<thead>
<tr>
<th>Freq. (GHz)</th>
<th>L</th>
<th>N</th>
<th>L₁</th>
<th>N₁</th>
</tr>
</thead>
<tbody>
<tr>
<td>140</td>
<td>420</td>
<td>164</td>
<td>400</td>
<td>80</td>
</tr>
<tr>
<td>165</td>
<td>420</td>
<td>64</td>
<td>400</td>
<td>30</td>
</tr>
<tr>
<td>220</td>
<td>260</td>
<td>104</td>
<td>240</td>
<td>50</td>
</tr>
<tr>
<td>300</td>
<td>190</td>
<td>84</td>
<td>170</td>
<td>40</td>
</tr>
</tbody>
</table>

Figure 3.21. Comparison of the reflection phase of the metasurface made of dogbone and the gangbuster at 220 and 300 GHz.

3.5 Conclusion

This chapter presents the use of a metasurface below the OCA (e.g., dipole) as an AMC reflector. Two different configurations, in terms of the location of the metasurface ground plane, are investigated. One is with a ground plane below the silicon and the other is having the ground plane at the lower metal layer.

With the ground plane below the silicon, we have obtained +1.2 dB simulated gain for a dipole antenna placed above a 4x5 dogbone array at 94 GHz. However, as the previous
discussion about OCA design concerns, a shielding ground plane at the lower metal layer is demanded for robust OCA radiation performance.

With the ground plane at the lower metal layer, it is discovered that the AMC property of the metasurface cannot be guaranteed because of the extremely-thin thickness between the periodic structure and the ground plane. When metasurfaces do not show AMC property, the metasurfaces cannot improve the gain of the antenna on the top. A transmission line model is proposed to understand better the underlying mechanism how to improve the metasurface design in order to obtain the AMC property. To prove the validity of the proposed lumped model, two different metasurface structures, a dogbone shaped structure [70, 71] and a mushroom structure [69, 86, 87] are studied respectively, assuming different level of losses in conductors. Based on the model, the threshold condition of AMC appearance and the bandwidth of AMC are proposed. A few guidelines to improve the performance of AMC are provided.
Chapter 4 Using a Metasurface as an On-chip Antenna Directly

4.1 Introduction

For a robust OCA design, a metallic layer to shield the antenna from the silicon substrate is desired. This can be achieved by placing a ground plane at the lowest metal layer M1, such that the extremely thin silicon dioxide layer acts as the antenna substrate. However, because of the extremely thin thickness of silicon dioxide (i.e., less than 1% of the wavelength in silicon dioxide), the antenna bandwidth is, in general, very narrow. In [88], an elliptical slot antenna with a ground plane at M1 shows a 3.9% impedance bandwidth around 94 GHz. In [89], a slot antenna at top metal layer backed with a cavity using M1 shows a 6 GHz (4.3%) bandwidth at 140 GHz according to full wave simulations, which is one the best reported results to date.

Instead of using the metasurface traditionally as a reflection plane [73, 87, 90], it has been conceptually shown in [91] that the metasurface can be used as a radiator itself. In this work, that concept is extended and further developed to design and implement an on-chip antenna with its accompanying matching network at mm-wave frequencies, including fabrication and experimental results. The design was implemented at 94 GHz using a 0.18 µm BiCMOS process.

The design of the novel fully on-chip metasurface antenna (without a dipole above it) is detailed including the description of the leaky mode inside the metasurface, which has been proven as the radiation mechanism of this antenna. Later, the design of an on-chip miniature size balun used to feed the proposed antenna will be presented. In the end, we present the experimental results including the input reflection, broadside gain, radiation patterns for the metasurface antenna, and the s-parameters for the balun.
4.2 Design of an On-chip Metasurface Antenna at 94 GHz

4.2.1 Mode Analysis

Without a dipole on the top, the dogbones in the proposed metasurface-antenna design are placed on M6, the most top metal layer with the ground plane at the bottom layer (M1) as shown in Figure 3.2 (b). Results will be further optimized using both M5 and M6 for the dogbones, as explained later in this paper. This results in an extremely thin thickness for the fully on-chip antenna equal to 0.27\% of the free space wavelength and around 0.5 \% (i.e., 1/200) of the wavelength in the dielectric, making this possibly the thinnest metasurface antenna ever fabricated.

The in-plane modal analysis in [91] shows that a metasurface made by a periodic array of dogbone shaped conductors over a ground plane is able to support a TM “improper” leaky mode in the $x$-direction at a frequency range close to the magnetic resonance. Recall that an improper leaky wave decays along the $x$-direction and grows exponentially in the $z$ direction. This is usually the typical forward wave used in a leaky wave antenna [91, 92]. Indeed, the study in [91] was focused on the anti-symmetric mode supported by an array of paired and tightly coupled dogbone conductors, as in [70]. However, because of symmetry, those results are analogous to the case studied here: an array of dogbones over a conductor. In the following, we show that the radiation mechanism of the on-chip metasurface antenna is analyzed in terms of a TM leaky wave excited in the metasurface, propagating along the $x$-direction, as shown in Figure 4.1. The full wave simulation is conducted using the finite-element-method provided by HFSS for six dogbone elements on M6 along the $x$-direction, assuming the periodic boundary condition along the $y$-direction. The dogbone element at the beginning of the row is excited by a lump port at its
edge (Figure 4.1). The dimensions of the dogbones are \( A = 0.43, B = 0.43, B2 = 0.12, B1 = 0.42, A1 = 0.12, \) and \( A2 = 0.42 \) (all in mm) according to Figure 3.1.

Figure 4.1. Six dogbone elements aligned in \( x \)-direction with period \( A \), and having periodic boundaries along \( y \), for high impedance surface design. The dashed line represents the exponential decay of a leaky wave excited at one edge.

The field over the “chain” of dogbones in Figure 4.1, excited at one end, is sampled 2 \( \mu \)m below the central bar of each dogbone, once per unit cell in the \( x \)-direction with period \( A \), at \( x = (n - 1) \times A + 150 \mu \)m, where \( n \) is the element number (in this case, \( A = 430 \mu \)m). The plot of the \( z \)-component of the electric field versus dogbone element number \( n \) is shown in Figure 4.2. As it can be seen, for the four frequencies specified in Figure 4.2 the field is decaying exponentially in the first several elements, as is typically the case for a leaky wave. As discussed in [93], the leaky mode is dominant in the first few dogbone elements close to the excitation, and away from...
it the so called “spatial field” term starts to play a major role, since the leaky wave itself becomes too weak. For more information about the excitation of leaky waves and spatial field terms in periodic structures, see [94, 95]. Results in Figure 4.2 show that the decay depends on the frequency. The phase variation of the field, evaluated at the same sample locations along the \(x\)-direction, is shown in Figure 4.3, for a few frequencies close to the one of interest (94 GHz). The phase variation is more or less linear with the element number.

![Figure 4.2. Field magnitude versus the element number (1~6) along the \(x\)-direction. Normalized with respect to the magnitude of the first sample.](image)

![Figure 4.3. Phase of fields versus element number (1~6) along the \(x\)-direction.](image)
Therefore, the field is highly dominated by a single leaky mode excited at the edge of the
dogbone chain in Figure 4.2. Accordingly, every electric field component at the sampled points
is expressed by

\[ E(x = nd) = E_0 e^{-j\beta_x nA} e^{-\alpha_x nA}. \]  

in which \( A \) is the period of the dogbones along \( x \), \( E_0 \) is the field at the sample point closest to the
excitation, and \( \alpha_x \) and \( \beta_x \) correspond to the attenuation and phase constants of the leaky mode.

By curve-fitting for the first four dogbone elements from the excitation (i.e., \( n=1,\ldots,4 \)), \( \alpha_x \) and \( \beta_x \)
are retrieved. Their values, normalized by the free space wavenumber \( k_0 \), are plotted in Figure
4.4. The phase constant of the leaky mode is smaller than that of the free space wavenumber
below 97 GHz, implying that this mode is in the fast-wave region and hence radiating. The
attenuation constant \( (\alpha_x) \) decreases as the frequency increases, whereas the phase constant \( (\beta_x) \)
shows the opposite trend. At 94 GHz, \( \beta_x \approx \alpha_x \), which has been indicated as the optimum condition
for leaky wave radiation in the broadside direction [96]. Note that this is not a “standard” leaky
wave antenna as the attenuation constant is not small, i.e., we are not exploiting leaky waves to
create a large radiating aperture to form a highly directive antenna. Indeed, in a highly directive
leaky wave antenna, the leaky mode would dominate the field on a very wide antenna aperture.
Instead, here we expect that the spatial wave may also contribute to the total radiation [93], and
the leaky mode is attenuating with \( \alpha_x \approx 0.6k_0 \), and therefore cannot cover a large aperture, though
it is still radiating because \( \beta_x \leq k_0 \).
According to the full wave simulation, the phase of the reflection of a plane wave normally incident over the metasurface does not exhibit zero-phase crossing. However, the magnetic resonance still appears, associated with currents flowing in opposite directions on the dogbone and the ground plane [70, 97]. The absence of the zero-phase reflection is attributed to ohmic losses.

It is known that close to the magnetic resonant frequency, there are strong ohmic losses on the dogbone due to the resonant current excited on the central part of dogbone and on the ground plane. Indeed, considering the dogbone over the ground plane, and assuming a current $I$ flowing on the central part (assumed a uniform current in the $x$ direction, for simplicity) connecting the two capacitors at the ends, the total ohmic loss in the dogbone conductor is $\frac{1}{2} I^2 R_{eff}$ where the resistance of the dogbone is approximately given by $R_{eff} \approx \sigma_s (A_2 - A_1) / B_2$. An analogous resistance contribution associated with the current flowing over the ground plane should also be added. Therefore, to reduce the ohmic loss on the dogbone, a wider width of the
dogbone central bar (i.e., a larger B2) is desirable. However, a wider dogbone central bar implies also a smaller inductance since the inductance associated with a single dogbone’s unit cell, comprising the dogbone and the ground plane, is given by \( L_{\text{eff}} \approx \mu_0 h (A2 - A2) / B2 \). The magnetic resonant frequency is approximately estimated by \( \omega \approx 1 / \sqrt{L_{\text{eff}} \cdot C_{\text{eff}}} \), where \( C_{\text{eff}} \) is the effective capacitor representing the two capacitance effects at the edges of the dogbone shown in Figure 4.5. The effective capacitor is mainly contributed by (i) the capacitance between the two dogbone arms and the ground plane, and (ii) the capacitive coupling between two adjacent dogbones in the x-direction, when they are very close (gaps of the same order of h). A smaller resistance (by increasing \( B2 \)) implies a smaller inductance that will push the resonance to higher frequency. To compensate for this, in order to keep the same resonant frequency, the effective capacitor should be increased. For instance one could increase the size of the dogbone “arms” \((A1B2)\). However, it would be preferable to keep a small size dogbone. Therefore, in this paper to increase the effective capacitance, we stack and combine two metal layers (M5 and M6) to form the dogbone arms such that the distance to the ground plane is smaller. In the layout, the dogbone arms at M5 and M6 are connected with array of 2 \( \mu \text{m} \times 2 \mu \text{m} \) vias with the period of 4 \( \mu \text{m} \) in both x and y axis. Furthermore, to increase the \( C_{\text{eff}} \), the dogbones are put close to each other in order to increase the capacitive coupling between the adjacent elements.
Figure 4.5. Unit cell of the metasurface: a dogbone element comprised of metals on M6, M5, and the vias in between. The equivalent lump model is indicated. Extra capacitive effect occurs towards adjacent elements in the x-direction.

Within a total area limit, the design rule of the single dogbone element dimension is based on keeping the magnetic resonance around 94 GHz, while choosing \( B_2 \) as large as possible to decrease losses. The final optimized dimensions for the dogbones, which occupy both M5 and M6 as in Figure 4.5, are \( B_2 = 250 \, \mu m \), \( B_1 = 420 \, \mu m \), \( A_1 = 100 \, \mu m \), \( A_2 = 500 \, \mu m \), \( A = 506 \, \mu m \), and \( B = 430 \mu m \).

### 4.2.2 Metasurface Antenna Design

Figure 4.6 shows the micrograph of the fabricated metasurface antenna made by a 3×4 array of dogbones with the unit cell dimensions provided above. As shown in Figure 4.6, the two center rows of the dogbone are connected with metals bars to enhance antenna gain according to full wave simulations. A differential twin line is used to feed the antenna at the end of the two center rows of dogbone array.
Figure 4.6. Micrograph of the metasurface antenna including matching network and balun.

Figure 4.7 shows the simulated broadside gain for the proposed metasurface antenna made of a 3×4 array of dogbones. The dogbone array is fed by a differential mode from a waveguide port in the HFSS simulation. Two cases are considered: an array of dogbones on two metal layers (M5 and M6 as shown in Figure 4.5), and the an array of dogbone placed only on M6. When using array of dogbones on two metal layers, the dimension of the dogbones are the ones specified at the end of Section 4.2.1. Comparing gain results for these two cases, it is observed that using two metal layers for each dogbone improves the gain by 2 dB at the center frequency (94 GHz), as expected simply because the overall loss is lowered in this case.
Figure 4.7. Comparison of the broadside gain versus frequency of the metasurface antennas when the dogbone is using (i) two metals layers (M5 and M6) (ii) single metal layer (M6 only).

For the purpose of the antenna measurement, a W-band balun is used to convert the single-ended signal associated with the W-band G-S-G probe to a balanced on-die differential signal. Also, an impedance matching network comprised of twin line and capacitors is designed between the metasurface antenna and balun (see Figure 4.6).

The final geometry and the size of the dogbone array on M6, including the balun is shown in Figure 4.6. The ground plane under the dogbone array on M1 at each edge is 100 µm and 150 µm larger than the dogbone array in x- and y-direction, respectively, making an area of 2.2 mm × 1.6 mm. Due to the extremely thin thickness of this metasurface, surface waves travelling along the silicon dioxide substrate carry very little power and therefore other dimensions of the finite size ground plane underneath the M6 dogbone layer would have little effect on the antenna gain. Indeed, according to the full wave simulations, as long as the length and width of the ground plane on M1 is 30 µm larger than that of the dogbone array on M6 (to collect all the fringe capacitance effects), there is literally no change in the antenna performance.
with increased size of the ground plane. This attribute is considered to be advantageous compared to other OCA design which uses ground plane below the silicon substrate. Furthermore, it is important to note that a full ground plane on M1 provides shielding to the silicon-based substrate under M1.

To satisfy the metal density rule, an array of 5 µm × 5 µm dummy fills are placed on all the metallic part of the antenna, on both the ground plane on M1 and the dogbones on M5 and M6. The dummy fills are also applied to other components mentioned in this paper, including the balun and microstrip line. It should be noted that due to the large memory requirements dummy holes and vias have not been included in the full wave simulations. For simulation purposes, the array of vias between M5 and M6 is replaced by bulk metal.

4.2.3 Effect of Metasurface Array Size on Antenna Gain and Directivity

This work also investigates the effect of the number of metasurface elements on the antenna gain. Simulations were conducted for metasurfaces with different numbers of array elements: 4×3 (the fabricated one), and 4×2, 6×2 and 6×3. For all these array sizes, the same twin line is used to feed the metasurface, and the ground plane on M1 is chosen as 100 µm larger than the dogbone array at each edge.

In Table 4.1, the broadside directivity and gain for four different metasurface dimensions are compared. It is observed that the larger array size results in higher the antenna gain and directivity, although the amount of gain increase could be very small.
Table 4.1. Simulated peak gain at 94 GHz for the metasurface antenna with different array sizes.

<table>
<thead>
<tr>
<th>Size</th>
<th>4x3</th>
<th>4x2</th>
<th>6x2</th>
<th>6x3</th>
</tr>
</thead>
<tbody>
<tr>
<td>Directivity</td>
<td>7.6 dB</td>
<td>7.2 dB</td>
<td>7.9 dB</td>
<td>8.3 dB</td>
</tr>
<tr>
<td>Gain</td>
<td>-1.8 dB</td>
<td>-3 dB</td>
<td>-2.78 dB</td>
<td>-1.5 dB</td>
</tr>
</tbody>
</table>

Figure 4.8 shows the broadside gain versus frequencies for different array sizes. It is interesting to see that the peak of the broadside gain always occurs around 94 GHz, despite the array dimension, since the dogbone elements are kept identical in all cases. This is an important consideration in the design. Indeed, this is explained by noticing that the magnetic resonance frequency of the dogbone metasurface is very close to the leaky mode frequency in the metasurface as shown in [91]. Furthermore, one should note that the leaky mode is responsible for the antenna radiation, especially considering the optimum condition in Figure 4.4. This represents simplicity in the design: the peak gain frequency of the proposed antenna could be estimated accurately by investigating the resonance frequency of a single element, which could be simply extracted by full wave simulations using periodic boundary conditions.

An array size of 4x3 was chosen for the fabrication, which accounts for a trade-off between the antenna gain and total area. It is noteworthy that decreasing the array size to 4x2, saving almost one-third of the die area compared to the 4x3 array, only lowers the antenna gain for 1 dB, which could also be considered in future designs. These simulated gains are among the highest values obtained for the fully OCAs with a full ground plane to shield the antenna from the substrate. For an OCA design that is constrained to a smaller chip area, a smaller array size of 4x2 dogbones also provides a reasonable compromised solution.
4.2.4 Miniature Sized Marchand Balun

The proposed antenna requires a differential feeding. To obtain a differential signal from a single-ended probe, a balun is to be designed. A Marchand balun [98] composed of two quarter wavelength couple lines is designed and developed for this design. The Marchand balun has been used for wide bandwidth designs at RF and microwave frequencies [99]. In [100], the Marchand balun was also implemented in (Bi)CMOS technology at 60 GHz. Different techniques to miniaturize the dimension of the Marchand balun have been investigated in [101, 102].

Figure 4.9 shows the top view of the Marchand balun designed using meander shaped couple line, referring the idea in [101]. The Meander-shape couple line replaces straight couple line and greatly shrinks the profile of the balun. Both the top half and bottom half of the couple lines are in the length of a quarter wavelength. The gap between the coupled lines is 2 µm. The mitered corners are used for bending the couple line while minimizing the insertion loss at the bends. The locations of the two differential outputs are shifted a certain distance away from the
center to achieve a near-perfect 180 degree out of phase difference between the two outputs. The total area of the balun is $140 \, \mu m \times 200 \, \mu m$.

Figure 4.10 shows the simulated input reflection and transmission to two output modes, one differential mode and one common mode of the balun. It can be observed that the designed balun performs well in the whole W-band, from 70 GHz to 110 GHz, with below $-10 \, dB$ input reflection, lower than $-20 \, dB$ transmission to common mode output. The average insertion loss of the balun over W-band is around 1.5 dB. The balun is characterized experimentally in the next section.

Figure 4.9. Dimension of the marchand balun.
Figure 4.10. Simulated input reflection and transmission to the differential mode and the common mode of the designed balun.

4.2.5 Measurement Results

For on-chip antenna measurement, one of the detrimental factors which will degrade the measurement accuracy is the possible electromagnetic interference (EMI) between the measuring probe and the antenna due to the proximity between them, as shown in [48]. To alleviate possible EMI for on-chip antenna measurement, a back-feeding scheme could be used for in-package antenna measurement as in [62]. However, due to the difficulty in having TSVs in commercial (Bi)CMOS processes, it is not straightforward to feed the OCA from the back-side of the radiating chip.

To mitigate this issue, we consider separating the antenna and the measuring probe with a rather long on-chip feed line. In the present characterization, we have used a 2.5 mm long microstrip line between the probe pad and the input of the balun. To accurately characterize the
insertion loss, another 2.5 mm long microstrip line is fabricated on the same die, and its measured and simulated results are shown in the Appendix B, leading to an insertion loss of 3dB.

To characterize the performance of the balun using single-ended probes, a back-to-back configuration made of two identical baluns is fabricated and the micrograph is shown in Figure 4.11.

Figure 4.12 shows the simulated and measured input reflection and transmission of the two back-to-back baluns. It can be observed that the measured results match well with the simulated ones, except for approximately 1dB higher insertion loss. This result is consistent with the long microstrip line measurement shown in Appendix B. Considering the symmetry, the measured insertion loss of a single balun is around 1.5 dB.

Figure 4.11. Micrograph of two same baluns in back-to-back configuration.
Figure 4.12. Comparison of simulated and measured input reflection and transmission of the back-to-back baluns in Figure 4.11.

The differential input impedance of the metasurface before the feed twin line is shown in Figure 4.13(a). Without any matching circuit, the metasurface shows a relatively flat input resistance and reactance in the frequency range between 90 GHz to 102 GHz, which implies a wide input bandwidth after proper matching. The matching circuit is shown in Figure 4.13 (b). The antenna matching network comprises a 230 µm long twin line ($Z_{0,\text{diff}} = 85 \, \Omega$) and two finger capacitors (22 fF) in series with each line. When the matching network is connected in between the metasurface and the balun, the parameters of the matching network were optimized to achieve a largest possible impedance bandwidth at the input of the balun.
The antenna input reflection is measured at the input of a 2.5 mm long microstrip line. Figure 4.14 shows the comparison between the simulated and measured antenna input reflection ($S_{11}$) from three separate antenna samples at the input of the long microstrip line. The measured results for the three samples are consistent. The simulated $-10$ dB input bandwidth covers a wide frequency range between 90 and 102.5 GHz. The measured $S_{11}$ shows a certain discrepancy with the simulated one, especially in the frequency range between 95 to 110 GHz. This could be due to several reasons, including:
(i) The full wave simulation does not include the dummy holes or the vias between M5 and M6. This affects the current flowing along the dogbone and the ground plane. If one considers the current flowing on the dogbone arms, because of the holes and the vias, the current flowing on the top surface of M6 could partially travel into the surfaces below, including the bottom surface of M6 and even the top and bottom surface of the M5. The metal layers M5, M6 and the vias in between shares similar thickness around 2 μm. Although the effect could be trivial in a short distance, the total effect could be summed up, and could thus play a strong effect since the antenna has a length (x-direction) longer than one $\lambda_g$ (the guided wavelength of microstrip line on silicon dioxide substrate).

(ii) Possible variations in the thickness of metal layers and dielectric layers as stressed in [88]. Considering the thickness between M5 and the ground plane on M1 is less than 6 μm, a possible variation of +/- 0.5 μm in any layer thickness would result in a large deviation in the antenna substrate thickness.

(iii) Variations in the material properties (dielectric permittivity and conductor conductivity) at mm-wave frequencies compared to microwave frequencies.

Nevertheless, since the measured $S_{11}$ between 90~102.5 GHz are all below or close to –10 dB, the deviation between simulation and measurement are reasonably small from the quantity point of view. To mitigate the deviation in the future work, it is desired to perform numerical simulations with fine details and have more accurate material electrical properties.
The same equipment setup and calibration scheme shown in Figure 4.15 (a), same as what is described in [48], is used to measure the radiation property of the metasurface antenna. The OCA is measured in the receiving mode by receiving radiation from the transmitting W-band calibrated horn antenna which is fed by a frequency multiplier. The antenna gain is calculated as

\[ G_{AUT} = P_{AUT} - P_{horn} + G_{horn} + Loss_p. \]  

in which all the terms given are in dB scale: \( P_{AUT} \) and \( P_{horn} \) are the power received at the spectrum analyzer when using the antenna under test (AUT) and an additional calibrated horn as receiving antennas, respectively with gain \( G_{horn} \) (24 dBi at 94 GHz). Furthermore, \( Loss_p \) is the total insertion loss including that from the probe (1dB), the balun (1.5dB), the 2.5 mm long microstrip feed line (3dB) and an estimated transition loss between probe tip and the on-chip CPW (0.5 dB), with a total loss of 6 dB. To suppress the reflection due to the presence of
metallic objects close to the antenna, thin mm-wave absorbers are used to surround the antenna, and also to cover the wafer chuck and probe-head as shown in Figure 4.15 (a) (b).
In Figure 4.16, the measured broadside gain, indicated by dashed line, is obtained by averaging the measured data from three distinct gain measurements. The measured results show good correlation in the 3 dB gain bandwidth of the antenna with the full wave simulation. The peak gain is around –2.5 dBi in the frequency range between 92 to 95 GHz. The 3 dB gain bandwidth is around 8 GHz, from 89 GHz to 97 GHz. To the author’s knowledge, this is probably one of the highest broadside gains and one of the largest gain bandwidths achieved to date at mm-wave frequencies by an extremely thin fully on-chip antenna, with a solid ground plane on M1. Considering the difficulties in measuring the fields on the metallic probe station and also considering the variety of calibration required for the measurement, the accuracy of the measurement is estimated as ±1~1.5 dB.
Figure 4.16. Comparison of measured and simulated antenna broadside gain versus frequency.

Table 4.2 shows a comprehensive list of recent published work [88, 89, 103-110] on single-feed mm-wave and Terahertz (THz) CMOS and BiCMOS fully on-chip antennas, radiating at the broadside direction, with a ground plane at the lower metal layer. Therefore for all these antennas, the substrate is extremely thin and the radiating zone is shielded from the lossy silicon substrate. Compared with other work, our proposed antenna shows the highest gain at W-band, and the widest relative impedance bandwidth achieved so far among all the shown designs at mm-wave and THz. It should be mentioned that when the frequency increases, and hence the substrate becomes electrically thicker, our proposed antenna has the potential to achieve even wider relative bandwidth and gain. Similarly, the improvement in terms of gain and bandwidth can be obtained also at W-band when considering other fabrication processes that allow substrate thickness larger than the one considered here (10 μm).

The normalized measured radiation patterns of the antenna at three different frequencies, including 94, 91.5 and 96.5 GHz in both E- and H-plane are shown in Figure 4.17. Each
measured pattern is compared against a simulated one. Due to the presence of certain obstacles in the probe station, the pattern measurement cannot be carried out in the complete E- and H-planes (see also [48]). In the H-plane, the measurement was done over only one quadrant and the range is limited by 40 degrees from the broadside direction. In the E-plane, the limit was 12 degrees in one quadrant and 50 degrees in the other quadrant. The pattern measurement was repeated twice in each plane and the curves shown are based on the averaging these two measured results. Measured patterns and simulated ones have similar trends. The ripple of the measured radiation pattern is possibly due to the interference between the probe and the AUT during measurement (see also [48]).
Table 4.2. Performance comparison of (Bi)CMOS based OCA over an extremely-thin substrate with a ground plane at the lower metal layer.

<table>
<thead>
<tr>
<th>Reference</th>
<th>Process</th>
<th>Antenna type</th>
<th>Antenna substrate thickness (µm)</th>
<th>Freq. (GHz)</th>
<th>−10dB impedance BW (GHz)</th>
<th>Relative impedance BW (%)</th>
<th>Gain (dBi)</th>
<th>Size (mm×mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>This work</td>
<td>180 nm BiCMOS</td>
<td>Leaky wave</td>
<td>10</td>
<td>94</td>
<td>10</td>
<td>10.6%</td>
<td>−2.5</td>
<td>2 × 1.3</td>
</tr>
<tr>
<td>[88]</td>
<td>130 nm CMOS</td>
<td>Elliptical slot</td>
<td>11</td>
<td>90</td>
<td>2.5</td>
<td>2.8</td>
<td>−6</td>
<td>1.7 × 1.1</td>
</tr>
<tr>
<td>[89]</td>
<td>180 nm BiCMOS</td>
<td>Slot</td>
<td>10</td>
<td>140</td>
<td>5</td>
<td>3.6</td>
<td>−2</td>
<td>1.2 × 0.6</td>
</tr>
<tr>
<td>[103]</td>
<td>130 nm CMOS</td>
<td>Slot</td>
<td>NA</td>
<td>9</td>
<td>NA</td>
<td>NA</td>
<td>−10</td>
<td>0.5 × 0.5</td>
</tr>
<tr>
<td>[104]</td>
<td>130 nm CMOS</td>
<td>Patch</td>
<td>NA</td>
<td>60</td>
<td>0.81</td>
<td>1.3%</td>
<td>−3.3</td>
<td>1.58 × 1.22</td>
</tr>
<tr>
<td>[105]</td>
<td>180 nm BiCMOS</td>
<td>Slot</td>
<td>10</td>
<td>360</td>
<td>NA</td>
<td>NA</td>
<td>−2.2</td>
<td>0.25 × 0.2</td>
</tr>
<tr>
<td>[106]</td>
<td>250 nm BiCMOS</td>
<td>Patch</td>
<td>15</td>
<td>79</td>
<td>4</td>
<td>5%</td>
<td>−1.3</td>
<td>0.95 × 0.85</td>
</tr>
<tr>
<td>[107]</td>
<td>130 nm BiCMOS</td>
<td>SIW</td>
<td>NA</td>
<td>410</td>
<td>12</td>
<td>2.9%</td>
<td>−0.5</td>
<td>0.5 × 0.2</td>
</tr>
<tr>
<td>[108]</td>
<td>45 nm CMOS</td>
<td>Patch</td>
<td>4</td>
<td>410</td>
<td>NA</td>
<td>NA</td>
<td>−1.5</td>
<td>0.2 × 0.2</td>
</tr>
<tr>
<td>[109]</td>
<td>65 nm CMOS</td>
<td>Patch</td>
<td>NA</td>
<td>1000</td>
<td>24</td>
<td>2.4%</td>
<td>+1.5</td>
<td>0.068×0.045</td>
</tr>
<tr>
<td>[110]</td>
<td>180 nm CMOS</td>
<td>Patch</td>
<td>&lt; 10</td>
<td>60</td>
<td>1.5</td>
<td>2.5</td>
<td>−14.5</td>
<td>1.15 × 1.15</td>
</tr>
</tbody>
</table>
Figure 4.17. Measured and simulated antenna radiation patterns in both the E- and H-planes at 94, 91.5 and 96.5 GHz.

4.3 Design of Metasurface OCA at Terahertz Frequencies

For the OCA design at THz frequencies, since the insulator layer is becoming electrically thicker, the ohmic losses could be alleviated in the case when the ground plane is on one of the lower metal layers. With a ground plane on the lower metal layer, different types of antenna have
been implemented on (Bi)CMOS chip at THz frequencies, including slot antenna [105], substrate integrated waveguide antenna [107], and rectangular patch antenna [35, 109]. While moderate gain performance (−2 dBi ~ +2 dBi) has been achieved from these designs, the relative bandwidth of these antennas are very narrow (around or less than 2% of the center frequency). At THz frequencies, a wideband antenna is demanded for a wideband operation.

In this work, a similar design centered at 312 GHz, based on a periodic structure made by dogbone shaped elements, is presented. The geometric parameters of single dogbone element used in this paper are shown in Figure 3.1. Figure 4.18 shows the configuration of the antenna made by 3×4 dogbone arrays at the top copper layer (M9). The antenna ground is placed on M1. The thickness between M1 and M9 is 4.6 µm. Each three dogbones in the same row are connected in order to improve the antenna gain a little bit, according to the full wave simulation. The final optimized dimensions for the dogbones are B2 = 80 µm, B1 = 160 µm, A1= 30 µm, A2 = 160 µm, A = 166 µm, and B = 170µm. Similar to the discussion in Section 4.2.1, the dogbone elements are arranged close together in the x direction, in order to take advantage of the capacitive coupling between the adjacent elements (i.e., increase the C in the LC resonance condition) such that the center dogbone width could be widen a bit (thereby L is decreased to maintain the same resonant frequency). With a wider dogbone (larger B2), the antenna gain could be improved because of the reduced ohmic losses.
Figure 4.18. Leaky wave antenna composed by a 3×4 array of dogbone elements. The dogbones in the same row are connected to increase the antenna gain.

Similar to the analysis in Section 4.2.1, a row of dogbone elements with a periodic boundary condition is excited at the edge of the first element in the row, to extract the leaky wave propagation property. The dimensions of the dogbone elements follow that used in the antenna design. By curve-fitting for the first four dogbone elements from the excitation, \( \alpha_x \) and \( \beta_x \) correspond to the attenuation and phase constants of the leaky mode are retrieved. Their values, normalized by the free space wavenumber \( k_0 \), are plotted in Figure 4.19. The phase constant of the leaky mode is smaller than that of the free space wavenumber below 326 GHz, implying that this mode is in the fast-wave region and hence radiating. The attenuation constant \( (\alpha_x) \) decreases as the frequency increases, whereas the phase constant \( (\beta_x) \) shows the opposite trend above 305 GHz. At 315 GHz, \( \beta_x \approx \alpha_x \), which has been indicated as the optimum condition for leaky wave radiation in the broadside direction [96].
Figure 4.19. Attenuation ($\alpha_x$) and phase constant ($\beta_x$) of the leaky wave along the $x$-direction, which is dominant in the first 4 dogbone elements. They are normalized with respect to the free space wave number.

Figure 4.20 shows the simulated broadside gain versus frequency for the proposed antenna. The peak gain is $+1.5$ dBi at 312 GHz, which is the very close to the optimum broadside radiation frequency (315 GHz) as expected from the dispersion relationship in Figure 4.19. The 3dB gain bandwidth is 25 GHz, between 298 GHz to 323 GHz.

Figure 4.20. Simulated broadside gain versus frequency (with the antenna input as indicated by black dashed line in Figure 4.18).
Figure 4.21 shows the input differential impedance seen at the plane as indicated by the black dashed line in Figure 4.18 when the array is feed by a pair of 2 µm wide twin lines with a spacing of 8µm. The input impedance between 300 and 330 GHz shows a relatively flat property, which implies a potential for wideband input matching. The antenna input is connected to the output of a frequency tripler. Instead of matching the antenna input impedance and the frequency tripler output impedance to 50 Ω respectively, the co-design of the antenna and the frequency tripler was conducted by conjugate matching the antenna input impedance and tripler output impedance directly, in order to simplify the matching network and meanwhile achieve the wideband matching as described in [11].

![Figure 4.21. Input differential impedance seen at the plane as indicated in Figure 4.18 by the black dashed line.](image)

Figure 4.22 shows the simulated normalized radiation patterns in E- and H-planes at 312 GHz and also at the edge frequencies of the 3dBi gain bandwidth, which are 298 GHz and 323 GHz. The radiation patterns sustain a good shape among the 3 dBi gain bandwidth frequency range.
Figure 4.22. Simulated antenna radiation patterns (normalized) at 312, 298, and 323 GHz in (a) E-plane and (b) H-plane.
4.4 Summary

The novel concept of using the metasurface as the radiator directly without a dipole on the top is investigated. Inspired by the concept of high impedance surface (HIS), this metasurface is not used as a reflector below an antenna as commonly done. Instead, it is used as a radiator by itself. The extremely thin metasurface is composed of a patterned top two metal layers and the ground plane placed in the lowest metal layer in the process. The ground plane on the lowest metal layer of the process provides a solid shielding from the substrate and other possible circuitries. The fundamental of the antenna radiation and design are described.

The metasurface could radiate because of the leaky mode the periodic structure supports. A novel fully on-chip antenna based on a metasurface fabricated in a 180 nm BiCMOS process is presented. With a ground plane at the lower metal layer, although the antenna substrate is extremely thin, the measured antenna shows –2.5 dBi peak broadside gain with 8 GHz 3-dB gain bandwidth and an impedance bandwidth larger than 10 GHz. Compared to the other works in the literature in the same frequency bands, the mm-wave designs show the widest impedance bandwidth among the OCA designs shown so far in the literature, in which a ground plane at the lower metal layer is considered.

The design is also investigated at THz frequencies (at 310 GHz) in a 65 nm CMOS process. The antenna shows +1.5 dBi simulated gain at 312 GHz.
Chapter 5  On-chip Slow Wave Transmission Line

5.1 Introduction

Millimeter wave (mm-wave) circuits in (Bi)CMOS process are becoming feasible thanks to the process scaling. Potential applications include multi-Gbps high data-rate wireless communication, anti-collision radar and imaging systems. At mm-wave frequencies, wavelength is comparable to the dimension of on-chip components. Unlike what happens for transistors, scaling of silicon technology does not guarantee the performance enhancement of passive transmission line (TL) such as microstrip and co-planar waveguide (CPW). Accurate simulation and modeling are essential to characterize the on-chip passive components.

Slow wave TL (SWTL) is attracting since they require a shorter physical length compared with the common TL, especially in a tight die size budget condition. From [111], a typical structure of a SWTL with a slotted ground shield is shown in Figure 5.1. The extracted $L$, $C$ per unit length in the transmission line RLGC model based on the full wave simulation results are shown in Figure 5.2. The inductance per unit length of the SWTL is identical to that of the common CPW while the capacitance is larger than that of the common CPW. The slow wave property is achieved by increasing the distributed capacitance, locally, from capacitive coupling, e.g., adding floating metals below the TL. It was also shown in [111] that this type of SWTLs often exhibits lower losses per wavelength, but not always per unit length compared with the non-slow-wave counterparts.
Figure 5.1. (a) SWTL with a slotted ground shield in [111], (b) Grounded CPW, (c) common CPW line.

Figure 5.2. Simulated $L$ and $C$ of the RLGC parameters of a slow wave transmission line, a grounded CPW and a common CPW as reported in [111].
In this work, alternative solutions are presented to achieve slow wave propagation resorting also to an extra distributed inductance generated by magnetic coupling to split ring resonators (SRRs) in addition to the capacitive coupling. Two different designs are presented. One is with a vertical split ring resonators (SRRs) all under the signal trace and the other is having two horizontal back-to-back SRR between the CPW and the bottom ground plane.

5.2 Slow Wave CPW with Single Split Ring Resonator

5.2.1 Design and Analysis Method

Figure 5.3 (a) shows the proposed slow wave grounded CPW in the $yz$ and $xz$ planes, while the wave propagation is in the $y$ direction. Figure 5.3 (b) shows the cross-section view of the CMOS process considered. The CPW signal trace is on the top metal layer (M6) while the CPW ground plane is on both M1 and M6. The SRR, aligned with the CPW signal trace in the $y$-direction, is made up by metals on metal layers M2, M5 and vias in between. The open gap of the SRR is on M2. The SRR has the same width as the signal trace of the CPW. The magnetic field generated by the CPW signal trace couples to the SRR exciting a current thereby.

Two different methods are considered to retrieve the propagation property of the transmission line. The first one is based on the eigen-mode solver in HFSS, to solve the complex resonant frequency of the system and successively obtain the attenuation constant ($\alpha$), phase constant ($\beta$) of the modal wavenumber $\gamma$ ($\gamma = \alpha + j \beta$), according to the relation in [112]. Another method is to calculate the propagation constant based on the two port scattering parameters obtained by full wave simulations. The scattering parameters are transformed to its ABCD matrix form where $\gamma$ and the characteristic impedance $Z_0$ are extracted directly. Table 5.1 shows
comparison of the performance parameters obtained by the two methods mentioned above for the same slow wave TL design with SRR. The results show good match with each other.

Figure 5.3. Cross-section views of the slow wave transmission line with SRR supporting mode propagating in the y-direction. (a) Cross-section view in y-z plane of 2 SRRs below the grounded CPW, (b) stackup of the BiCMOS process, (c) cross-section view in x-z plane of SRR below grounded CPW.
Table 5.1. Comparison of the extracted propagation constant extracted by two different methods.

<table>
<thead>
<tr>
<th>Methods</th>
<th>$\lambda$ (mm)</th>
<th>$\alpha$ (mm$^{-1}$)</th>
<th>$\beta$ (rad · mm$^{-1}$)</th>
<th>$\beta/\alpha$ (rad)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Scattering-parameter</td>
<td>1.0404</td>
<td>6.039</td>
<td>0.1464</td>
<td>41.2</td>
</tr>
<tr>
<td>Eigen-mode</td>
<td>1.0396</td>
<td>6.044</td>
<td>0.137</td>
<td>44.1</td>
</tr>
</tbody>
</table>

5.2.2 Analysis and Simulation Results

Figure 5.4 shows the performance of the slow wave CPW with SRR of different lengths including 220 µm, 240 µm and 260 µm and common CPW obtained by the scattering parameters method. The other dimensional parameters are (all in µm): SRR open gap = 20, SRR width = CPW signal trace width = 12, CPW pitch = 32, and distance between two contiguous SRRs = 40. The relative dielectric constant of the substrate is 4.2.

It can be observed in Figure 5.4 (b) that with the additional SRR below the CPW signal trace, the phase velocity could be significantly controlled, compared to a common CPW, in particular it can be slowed down. Larger phase propagation constant, thus slower phase velocity implies a smaller physical length required to achieve a certain phase shift. The boost is particularly strong close to the resonant frequencies of the SRRs. SRR with different lengths exhibits different resonant frequencies and this provides a control parameter for the SWTL.

Because of the very subwavelength thickness, significant losses are introduced in the proposed SWTL at the resonance of the SRR ($f_0$). Figure 5.4(a) shows that at $f_0$, the attenuation constant $\alpha$ reaches its peak value (e.g., for the SRR with length of 260 µm, the resonance is
around 95 GHz). Meanwhile, $\beta$ reaches its peak value at a frequency $f'_0$ slightly lower than $f_0$ (e.g., 90 GHz for SRR in the length of 260 µm). Since losses are strong close at the resonance frequency, it is convenient to use the proposed SWTL at the frequency range slightly away from the resonance such that a moderate $\beta$ enhancement could be achieved while the attenuation is manageable (e.g., between 60~70 GHz, based on the designs shown in Figure 5.4).

Figure 5.4 (c) and (d) show the distributed serial inductance and shunt capacitance per unit length of the proposed slow wave CPW compared with a common CPW, based on the $RLGC$ lumped model. It can be observed that below the frequency $f'_0$, the proposed CPW shows an enhanced inductance, while this phenomenon disappears at the frequency above $f'_0$. This effect is solely due to the magnetic coupling between the SRR and the CPW. At the same time, the per-length capacitance is larger than that of a common CPW, which also contributing to slowing phase velocity. Note that the capacitance slightly decreases as the frequency increases towards $f'_0$, in contrast to the increase in the per-length inductance, providing a more or less constant increased phase velocity. By observing the direction of current flow in the CPW and in the SRR via full wave simulations, at the frequency below $f'_0$, the magnetic flux generated by the current on SRR has the same direction as that generated by the current flowing on the signal trace, thus reinforcing the magnetism through mutual inductive coupling. Therefore the equivalent distributed inductance is increased. In contrast, the current direction in the SRR is reversed at a frequency above $f'_0$ that implies that the overall inductance per unit length is smaller than that of the common CPW.
Figure 5.4. Comparison of the performance of the slow wave CPW with single SRR of different lengths with common CPW and the equivalent circuit model. (a) Attenuation and propagation constant, (b) serial inductance and shunt capacitance per unit length, (c) characteristic impedance and (d) circuit model of the slow wave TL with SRR.
The shunt capacitance per unit length of the proposed SWTL is always larger than that of common CPW due to the extra capacitance towards the bottom ground provided by the SRR metal. Therefore, the loading of the SRRs to the CPW provide both additional capacitance and inductance, thus allowing the characteristic impedance to be kept close to that of the common CPW (without modifying CPW dimensions). The characteristic impedance shown in Figure 5.4 (d) is more frequency-dependent than the common CPW, but nearly-flat near 60 GHz.

5.2.3 Circuit Model

Due to the relatively small electrical dimensions of the SRR the structure can be described by means of a lumped-element equivalent circuit in Figure 5.4 (f). $L_1$, $R_1$ and $C_3$ are the per-section inductance, resistance and capacitance of the CPW without SRR while the SRR is modelled by $L_2$, $R_2$, $R_3$ and $C_2$. The magnetic coupling and the electrical coupling between the CPW and the SRR is modelled by $M$ and $C_1$, respectively.

According to Figure 5.4, the comparison between the lumped circuit model and full wave simulations for the proposed design with SRR length of 240 µm shows good matching in terms of $\alpha$, $\beta$, $L$, $C$ and $Z_c$. The values for $R_1$, $L_1$, and $C_3$ was obtained based on RLGC model of the common CPW whereas $R_2$, $R_3$, $L_2$, $C_1$ and $C_2$ were calculated based on the physical dimension of the SRR using static electro- and magnetic models. These values were further optimized in a schematic circuit simulator (i.e., AWR) taking into account the mutual inductance ($M$). For the curve shown in Figure 5.4, the values of lumped elements representing a SWTL (with SRR length = 240 µm) are: $R_1 = 0.36 \, \Omega$, $L_1 = 0.041 \, \text{nH}$, $C_3 = 2 \, \text{fF}$, $R_2 = 0.44 \, \text{mQ}$, $R_3 = 2.16 \, \Omega$, $L_2 = 0.056 \, \text{nH}$, $C_1 = 30.6 \, \text{fF}$, $C_2 = 84 \, \text{fF}$, and $M = 0.034 \, \text{nH}$. 
5.2.4 Loss Compensation with Active Components

Additionally, losses introduced by the extremely flat SRR loading could be compensated by active elements composed by transistors which could in principle be used also for tuning. A good example is a non-foster circuit (e.g, negative impedance converter (NIC) as in [113]) that provides an equivalent negative resistor for loss compensation. For example, we consider a negative resistor value of $\text{–}600 \, \Omega$, in the range that could be typically realized at mm-wave using cross-coupled pair circuit topology in BiCMOS as shown in [114]. Figure 5.4 (a)-(c) also shows the SWTL parameters $\alpha$, $\beta$, $L$, $C$ and $Z_c$ with loss-compensation below 100 GHz by adding the ideal negative resistance in the equivalent circuit as indicated by the dashed lines in Figure 5.4 (d), which models the effect of attaching active circuits between the open gap at the bottom of the SRR.

5.3 Slow Wave CPW with Double Split Ring Resonators

Besides single SRR fully below the signal trace, another SWTL design with double split ring resonators as shown in Figure 5.5 was studied. Figure 5.5 (a) shows the top view of the slow wave grounded CPW propagating in the $x$ direction, based on same CMOS stackup as in Figure 5.3(b). The CPW signal trace is on the top metal layer (M6) while the CPW ground plane is on both M1 and M6. Figure 5.5 (b) shows the dimensional parameters of the double SRR structures, which are placed on M5 (indicated in blue color). The magnetic field generated by the CPW signal trace couples to the central bar of the SRR, which is below the CPW signal trace and thereby excites a current. The current flows around the split ring resonators and generates a
magnetic field along the \( z \) axis, which coincides with the direction of the magnetic field originated from the CPW signal trace. In the previous design of SWTL with single SRR fully below the signal trace, the magnetic coupling between the SRR and the CPW signal trace is limited by the thickness of insulator layers. When the SRR becomes thinner, the magnetic flux passing through the SRR also gets weaker. Therefore in the previous case, the only parameter to control the coupling is the length of SRR. Meanwhile, in this design with horizontal double SRR below the SRR, since the size of the SRR could be also adjusted by the width of SRR (as indicated in Figure 5.5), the magnetic coupling between the double SRR and the signal trace is only partially affected by the thickness of the insulator layer. One could adjust the pitch of the CPW while keeping the impedance, leaving more spacing between the signal trace and the ground of CPW and thereby increase the magnetic coupling between the signal trace and the double SRR.

Figure 5.6 shows the performance of the slow wave CPW with double SRRs of different lengths including 140 \( \mu \text{m} \), 170 \( \mu \text{m} \) and 200 \( \mu \text{m} \) and common CPW obtained by the scattering parameters method. The other dimensional parameters are (all in \( \mu \text{m} \)): SRR open gap = 20, Line width = CPW signal trace width = 12, Width = 62, CPW pitch = 32, and distance between two adjacent SRRs = 40.
Figure 5.5. (a) Top views of the slow wave transmission line with three double SRRs supporting mode propagating in the z-direction, (b) dimensional parameters of the double SRRs.

Compared with the result in Figure 5.4, similar phenomenon is shown in Figure 5.6, including a significantly boosted phase constant around the resonance, compared to a common CPW, and also significant increased losses. SWTLs with double SRRs of different lengths exhibit different resonant frequencies. The equivalent serial inductance per unit length shown in Figure 5.6 is relatively higher than that shown in Figure 5.4. It proves that the planar double SRRs provide a stronger magnetic coupling, compared with the single vertical SRR, to the CPW signal trace.
Comparison of the performance of the slow wave CPWs with double SRRs of different lengths with a common CPW, (a) attenuation constant, (b) propagation constant, (c) serial inductance, (d) shunt capacitance per unit length, (e) real part of characteristic impedance.

5.4 Summary

The novel designs of slow wave transmission line with split ring resonators completely below the trace line are proposed, leaving space free around the CPW. An equivalent circuit model and two different approaches to extract the propagation property of the transmission line are provided. Two different designs employing split ring resonators are studied. The designs
show possible control of both inductance and capacitance per unit section that in turns control both wavenumber and characteristic impedance. Extra losses can be compensated in principle using silicon based active components.
Chapter 6 Summary and Conclusions

As the final step of a fully-integrated transceiver system in silicon, on-chip antennas (OCAs) have attracted lots of interest. However, it is very challenging to achieve a high performance OCA at millimeter waves. The difficulties in designing high radiation efficiency antenna on CMOS chip are discussed. It is concluded that the radiation efficiency of OCA suffers either from strong dielectric losses in the silicon when the antenna ground is below the silicon or from the strong ohmic losses of the metals when the antenna ground plane is placed at one of the lower metal layers of the (Bi)CMOS process. Nevertheless, OCA with a ground plane at one of the lower metal layers is considered as a preferred solution since the presence of the metal ground plane is important to prevent wave leakage into the silicon substrate and render the antenna performance independent of the die size because of its shielding.

Several feasible millimeter wave on-chip antenna designs, suitable to be fabricated in CMOS technology without any additional process, are presented. With a ground plane at the lowest metal layer (M1), moderate antenna gain are obtained from the designs of a cavity-backed slot antenna, a substrate integrated waveguide slot antenna and an E-shaped patch antenna, all centered at 140 GHz. While these three designs show similar gain (around −2dBi), the E-shaped patch antenna shows the widest −10dB impedance bandwidth, which is around 10 GHz. Additionally, for the sake of wideband operation, a bowtie shaped slot antenna considering a ground plane below silicon is also presented. The standalone bowtie slot antenna, centered at 94 GHz, shows a measured gain of 0 dBi and an input bandwidth covering the whole W-band (from 70 to 110 GHz). The bowtie slot antenna was also implemented in a phased array configuration. Beam steering radiation patterns using on-chip antenna are successfully obtained.
Use of a metasurface in OCA design is investigated. We discuss on the fact that the artificial magnetic conductor (AMC) property cannot be guaranteed at the resonance when the thickness of a metasurface becomes very thin. An equivalent lumped circuit was proven successfully to model a metasurface including the effect of losses, especially suitable when the metasurface are electrically-thin such that the losses are mainly due to ohmic losses instead of the dielectric losses. Based on the equivalent model, we derive the threshold condition for AMC, and show that it may hold at certain incidence angle and not at others. We also derive a formula for the AMC bandwidth and find out that the bandwidth is directly related to the quality factors of the lumped model. Using these formulas, we obtain several AMC design rules, which could be useful to improve the bandwidth of AMC.

Instead of use a metasurface as a reflector, the concept of using the metasurface as an antenna directly is validated by both full wave simulations and measurement. It is known that for silicon based OCAs, the presence of ground plane on the bottom metal layer usually causes low efficiency and very narrow bandwidth. Despite that, the design of a fully on-chip antenna inspired by the studies on high impedance surfaces, with a full ground plane on the bottom metal layer, exhibits the great performance in terms of gain-bandwidth product at W-band for this class of antennas radiating at broadside. The design centered at 94 GHz shows a measured gain of –2dBi and an impedance bandwidth larger than 10 GHz. A similar metasurface antenna operating at Terahertz (THz) frequencies is also presented.

It is also true that with off-chip components, placed above the chip, the antenna gain could be improved, even significantly. Examples include using lenses [20], placing dielectric resonators [64, 115], or superstrate [88] on top of the chip and feeding those with an on-chip launcher. Though performance (efficiency and bandwidth) of these hybrid solutions are, in
general, better than that of fully on-chip antennas, these techniques require additional processes, which may increase the cost and design complexity. On the contrary, all the proposed fully on-chip antennas are designed in a silicon process with no additional post-fabrication processing. At sub-mm-wave and THz frequencies, it is expected that even better performance could be achieved by the proposed design due to the wavelength shrinkage.

The novel designs of slow wave transmission line with split ring resonators completely below the trace line is also proposed. The designs show possible control of both inductance and capacitance per unit section that in turns control both wavenumber and characteristic impedance.
Appendix A: Derivation of AMC Bandwidth

Using the RLC model, the reflection coefficient could be calculated as

\[
\Gamma = \frac{Z_{AMC} - Z_0}{Z_{AMC} + Z_0} = \frac{1 - \frac{Z_0}{R} \left(1 - \frac{\omega^2}{\omega_0^2}\right) + j \frac{\omega}{\omega_0} Q \left(1 - \frac{Z_0}{RQ^2}\right)}{1 + \frac{Z_0}{R} \left(1 - \frac{\omega^2}{\omega_0^2}\right) + j \frac{\omega}{\omega_0} Q \left(1 + \frac{Z_0}{RQ^2}\right)},
\]  

(A.1)

The bandwidth of an AMC is defined as the frequency range when the phase of \(\Gamma\) is between \(-90^\circ\) and \(90^\circ\). It is clear that the upper and lower limit of the bandwidth could be obtained by imposing the real part of \(\Gamma\) equal to zero. After rearranging (A.1), it can be noticed that finding the two limits is equivalent to solving the equation as below

\[
\omega^4 - \left(\frac{R^2}{Z_0^2} - 1 - \frac{1}{Q^4}\right)Q^2 + 2 \omega_0^2 \cdot \omega^2 + \omega_0^4 \left(1 - \frac{R^2}{Z_0^2}\right) = 0,
\]  

(A.2)

Here it is helpful to use \(Q_{c0} = Z_0 C \omega_0 = Z_0 / \eta\). Rearrange (A.2), we will have

\[
\omega^4 - \left(\frac{1}{Q_{c0}^2} - 1 + \frac{1}{Q^2}\right)\omega_0^2 \cdot \omega^2 + \omega_0^4 \left(1 - \frac{1}{Q_{c0}^2 Q^2}\right) = 0,
\]  

(A.3)

The solution of above equation is

\[
\omega_{1,2} = \omega_0 \sqrt{\frac{1}{Q_{c0}^2 - Q^2} + 2} \pm \sqrt{\frac{1}{Q_{c0}^4} + \frac{1}{Q^4} + \frac{2}{Q_{c0}^2 Q^2} + \frac{4}{Q_{c0}^2} - \frac{4}{Q^2}}
\]  

(A.4)

The bandwidth of the AMC is the difference between the two angular frequencies, which could be simplified as

\[
BW_{exact} = \omega_0 \sqrt{\frac{1}{Q_{c0}^2 - Q^2} + 2} - 2 \sqrt{\frac{1}{Q_{c0}^2 Q^2}},
\]  

(A.5)
Since $Q_{L0} = Q$, we have

$$BW_{exact} = \alpha_0 \sqrt{\left(\frac{1}{Q_{c0}^2} - \frac{1}{Q_{L0}^2} + 2\right) - 2 \sqrt{1 - \frac{1}{Q_{c0}^2 Q_{L0}^2}}}$$

(A.6)

**Appendix B: Measurement of a 2.5 mm Long Microstrip Line**

Figure B.1 shows the micrograph of a 2.5 mm long microstrip line with two ends connected to two bond pads. The bond pads are as those used in [48]. The parasitic capacitance of the bond pad is compensated with a shunt stub to achieve a stable impedance of 50 Ω at the input of the microstrip line over a wide frequency band (70 ~ 110 GHz).

![Micrograph of a 2.5 mm long 50 Ω microstrip line.](image)

Figure B.1. Micrograph of a 2.5 mm long 50 Ω microstrip line.

The simulated, including the bond pads, and measured input reflection and transmission of the 2.5 mm long microstrip line are compared in Figure B.2, providing a good agreement. For the insertion loss ($S_{21}$), there is an average 1dB ~ 1.8 dB deviation between the simulation and the measurement. The deviation stems from the fact that (i) the electrical properties of the materials at mm-wave frequencies are a bit different from those provided by the foundry (used in
the simulations) that are typically obtained below 10 GHz; and (ii) possible wave leakage into silicon substrate and variation to metal conductivity could exist due to the existence of the dummy holes, thereby creating slightly higher losses in the long microstrip line.

Figure B.2. Comparison of the simulated and the measured input reflection and transmission of the 2.5mm microstrip line in Figure B.1. The long transmission line has been used to feed the antenna.
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