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Wideband Microwave, Millimeter-Wave and Light-Wave Antennas

Hongyu Zhou
University of Colorado at Boulder, zhouhongyu2k@gmail.com

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WIDEBAND MICROWAVE, MILLIMETER-WAVE AND LIGHT-WAVE ANTENNAS

HONGYU ZHOU

B.S. Harbin Institute of Technology, China, 2007

M.S. University of Colorado at Boulder, 2009

A thesis submitted to the Faculty of the Graduate School of the University of Colorado at Boulder in partial fulfillment of the requirements for the degree of Doctor of Philosophy

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This thesis entitled:
Wideband Microwave, Millimeter-Wave and Light-Wave Antennas
written by Hongyu Zhou
has been approved for the
Department of Electrical, Computer and Energy Engineering

__________________________
Prof. Dejan S. Filipovic

__________________________
Dr. James McDonald

Date____________________

The final copy of this thesis has been examined by the signatories, and we
Find that both the content and the form meet acceptable presentation standards
Of scholarly work in the above mentioned discipline.
ABSTRACT

Wideband antennas are capable of maintaining consistent near- and far-field performance over wide bandwidths. With the rapid growth of information technologies and ever increasing needs for high data throughput, these antennas become increasingly important for modern communication systems. However, many challenges arise in the design of wideband antennas and their use in different regions of the frequency spectrum.

Majority of wideband antennas designed for commercial applications nowadays operate at microwave frequencies, for which low-cost and low-profile are often as important as antenna’s consistent electrical performance. This thesis first proposes a new feeding method to increase the bandwidth of the inherently narrowband patch antennas, the workhorse of modern communication industry. Compared with the previously published approaches, the proposed feeding technique delivers significantly increased impedance and far-field bandwidths, while maintains the antenna’s low-cost and low-profile properties. The associated challenges including radiation pattern degradation, mutual port couplings, and electrical sensitivity on structural variations are thoroughly discussed and carefully addressed.

Considering the congestion of lower microwave spectrum, modern wireless systems are often designed for millimeter-wave spectrum. Typically they require wideband millimeter-wave antennas capable of seamless, preferably monolithic integration with the system’s electronic
circuitry. Recent advances in micro-electro-mechanical technologies have contributed to the development of micromachining processes capable of achieving many desired features of millimeter-wave systems. The second part of the thesis demonstrates millimeter-wave log-periodic dipole array (LPDA) antennas designed for and fabricated with a thick photolithography manufacturing process. It is shown for the first time that millimeter-wave LPDA antennas can be reliably achieved in the millimeter-wave region. It is also demonstrated that it is possible to monolithically integrate different devices within the LPDA antenna without impacting the antenna performance. For low-cost dual-polarized wideband antenna solutions, millimeter-wave planar log-periodic antennas fabricated using printed circuit board (PCB) process are also investigated as an alternative to the LPDA antennas.

Finally, wideband on-chip optical antennas are demonstrated for low-loss low-latency optical interconnects for the next generation microprocessor multi-core systems. Due to the high metallic loss at these frequencies, silicon-on-insulator based dielectric antennas are developed and over 50THz bandwidth is demonstrated. Based on the designed optical antennas, wideband optical signal hubs for wavelength-division-multiplexed (WDM) channel interconnects and data broadcast are developed. Theoretical study shows the proposed interconnect solution provides significantly increased power efficiency compared with the traditional electrical interconnect solution.
DEDICATION

To my parents and my beautiful fiancée Lin.
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CHAPTER 1
INTRODUCTION

1.1 Background and Motivation

IEEE defines antenna bandwidth as “the range of frequencies within which the performance of the antenna, with respect to some characteristic, conforms to a specified standard” [1], indicating that antenna bandwidth is not solely defined by any predetermined constraints, instead, it is a concept highly related to specific cases. Most modern applications consider the antenna impedance match and the boresight gain as bandwidth criteria. The consistency of antenna far-field parameters, such as beamwidth, back-lobe level, cross-polarization level, axial ratio, etc., is also important for many applications and should also be considered.

If relatively constant performance is achieved for a specific criterion over a wide frequency range, the antenna is considered as wideband antenna. Due to the high signal throughput and broad frequency coverage, they are highly desired in modern communication and electronic warfare systems. However, even though wideband antennas have been studied for a century [2], numerous challenges still exist when their operation is desired in different regions of the frequency spectrum.

At microwave and lower frequencies, the operating principles and manufacturing techniques for the traditional wideband antennas have been well established. Typically microwave wideband antennas fall into three categories [3]. The first type introduces incoherent
resonance in the antenna structure, such as tapering the arms of the dipoles or monopoles, to achieve wide bandwidth. Examples include bowtie [4-5] and conical antennas [6-7]. The second type increases the antenna bandwidth by controlling field diffractions via careful geometry design or using absorptive loading, such as tapered-slot antennas [8], ridged horns [9] and resistively loaded dipoles [10]. The last type is the frequency independent antennas which employ frequency-scaled geometries to achieve ultra-wide bandwidth. Log-periodic antennas [11] and spiral antennas [12] are the best known antennas in this category. Some inherently narrowband antennas need further investigation on their designs to achieve wider bandwidth. One example is the microstrip patch antennas [13-19], which exhibit excellent properties for manufacturing and commercialization. However, due to its inherent narrow bandwidth, numerous emerging wideband and multi-band applications, such as wideband WLAN[19-20], mobile TV[21], LTE[22], WiMAX[23], etc., consider using multiple antennas to cover their whole bandwidth which significantly increases the manufacturing cost and lowers the overall antenna efficiencies due to the mutual coupling between antennas. A microstrip patch antenna that could fully support these bandwidths with a single broadband unit would significantly improve the performance.

In the millimeter-wave frequency region, wideband antenna development has been much slower due to the small feature size and insufficient manufacturing capabilities to reliably achieve these antennas in the last century. Recent advances in mechanical and material technologies bring new possibilities for the development of miniature millimeter-wave devices. Sequential surface micromachining technologies, such as E-FAB™ [24] and PolyStrata™ [25] make micron scale 3D millimeter-wave components achievable. Thinner and lower loss substrates and improved machining resolutions even allow the traditional printed circuit board
(PCB) manufacturing to be seriously considered. Thus, new millimeter-wave multi-octive wideband antennas compatible with the latest fabrication technologies have to be developed. The impact of the manufacturing restrictions of recently developed machining technologies on antenna performance needs to be investigated.

When contrasted to microwave and millimeter-wave antennas, antennas at optical wavelengths are newly sought-for devices. Using focused ion beam milling [26] or electron-beam lithography [27], the dimension accuracy of nanoscale optical antennas can be controlled within 10nm. Presently, optical antenna research mainly concentrates on near-field enhancing. They are primarily used for high-resolution microscopy and spectroscopy [28–29], optical sensors [30], photovoltaics [31], solid state lighting [32], and lasing [33]. The operating principles of these optical antennas are not directly related to their microwave counterpart. Instead, they are related to the optical plasmonic effects of metallic structures. The plasmonic resonance at optical frequencies can considerably shorten the antenna electrical dimension [34] and high material losses and fabrication limitations make the antenna feeding a huge challenge using traditional methods. However, emerging applications such as optical wireless on-chip interconnect [35] open a new avenue for optical antennas that can be fed easily and used effectively as signal transmitters and receivers.

Foreseeing the challenges discussed above, further developments in design and fabrication of wideband antennas in different frequency regions is increasingly important. This thesis provides comprehensive study of different types of wideband antennas in three spectral regions, i.e. microwave, millimeter-wave and light-wave. A new bandwidth enhancing approach is proposed for traditional narrowband microstrip patch antennas. Sequential surface micromachining and printed circuit board manufacturing processes are employed to fabricate
millimeter-wave log-periodic antennas. Dielectric material based optical antennas are developed on Silicon-On-Insulator (SOI) wafer with antenna link loss comprehensively measured in the infrared spectrum.

1.1.1 Microwave Wideband Patch Antennas

The concept of microstrip patch antennas was first introduced in 1953 [36]. Due to the lack of dielectric substrates with low loss tangent and good mechanical properties, practical patch antennas were not developed until the 1970s [33,37]. Since then, these antennas have been extensively studied due to their numerous advantages, such as light weight, low profile, low cost, conformability, and compatibility with integrated circuits.

However, patch antennas suffer from inherently narrow bandwidth. Depending on the substrate thickness and dielectric constant, the bandwidth of a traditional patch antenna typically ranges from 2% to 5%. This small bandwidth inhibits their use for many commercial and defense applications. To address this bottleneck, various bandwidth enhancement approaches have been proposed over the past few decades, including, but not limited to, planar multi-resonator patches [38-43], stacked patches [44-47], slot single layer patches [48-54] and capacitively probe fed patches [55-61].

For the planar multi-resonator approach, impedance bandwidths up to 25% can be obtained by placing several parasitic patches around the primary patch in the same layer, as shown in Fig. 1.1. However, this approach significantly increases the antenna physical size, making them unsuitable for many space constrained applications. Also, the radiation patterns are not stable throughout the impedance bandwidth. Another coupled resonator approach that uses stacked coupled patch structures, has been favored in most applications because of the size and
Fig. 1.1 Various gap-coupled multi-resonator patch antenna configurations. (a) Three patches gap-coupled along radiating edges. (b) Three patches gap-coupled along non-radiating edges. (c) five gap-coupled patches.

ease with which the coupling coefficient can be controlled by adjusting their separation height.

The geometries of two typical stack patch configurations are shown in Fig. 1.2. Impedance bandwidths up to 30% can be obtained by stacking two or more patches using multi-layer printed circuit board. Further bandwidth increase requires foam substrates [47], which significantly reduces the overall antenna robustness and increases the antenna design complexity. For slot single layer configuration, U-shaped slot [48-50] and E-shaped patch [52-53] are typically used

Fig. 1.2 Coaxially fed stacked patch antenna (left). Aperture coupled stacked patch antenna (right).
as shown in Fig. 1.3. The patch shape modification introduces a second resonance, which in turn increases the patch bandwidth to about 30%. However, due to the mode variation within the impedance bandwidth, high frequency radiation patterns exhibit significant distortion. Another method to obtain a second resonance is using capacitive probe feed [55-61]. Up to 40% impedance bandwidths can be obtained with consistent far-field performance by using a differential input [62]. The geometry of an L-probe fed patch antenna is shown in Fig. 1.4. However, foam material is required to support the patch above the ground plane, reducing mechanical robustness.

The majority of the patch bandwidth enhancement approaches proposed up to now are based on the above discussed methods. However, few of them are reasonable candidates for
practical applications due to a number of shortcomings. It is thus desired to develop a practically realizable method to further increase the patch bandwidth, while maintain good mechanical robustness for applications in diverse environments.

1.1.2 Millimeter-Wave Log-Periodic Antennas

Log-periodic antennas were introduced in the late 1950s by DuHammel and Isbell [63]. They are defined as structures whose electrical properties vary periodically with the logarithm of frequency. Shown in Fig. 1.5 are the two radio frequency log-periodic antenna realizations, log-periodic dipole array (LPDA) antenna and planar log-periodic antenna. The LPDA antenna consists of two in-plane sets of crisscrossed monopoles whose lengths and separations are related via the growth rate $\tau$. The alternating monopole configuration is required to achieve backward end-fire radiation in the direction of the small monopole pairs. The feeding point of the antenna is typically located at the antenna tip (high frequency end) to prevent the higher order mode radiation from the large dipoles when the antenna is operating at the higher in band frequencies. On the contrary, planar log-periodic antenna has all its monopole elements in the same plane.

![Fig. 1.5 Frequency independent LPDA antenna (left) and planar log-periodic antenna (right).](image-url)
with the feeding point laid in the center of the antenna structure. This construction makes the antenna bidirectional but easily machinable using PCB process. Planar log-periodic antennas are typically designed with two, four or eight arms for different operating modes and polarizations. For four arm planar log-periodic antennas, dual-linear or dual-circular polarizations can be obtained by implementing different phase progressions between the feeding points.

At VHF/UHF frequencies, an LPDA antenna can be easily fabricated using metallic tubes. It is fed by a coaxial cable brought from the low frequency end into one of the hollow booms with its outer conductor attached to the host boom and the inner conductor to the opposite boom tube at the high frequency end [64], as shown in Fig. 1.6. However, for higher microwave frequencies (typically up to 10GHz), this approach is not feasible due to the reduced antenna size and the PCB based replacements are widely used [65-69]. In this case, the feeding coaxial line is either attached to the board with its outer conductor touching one of the booms and center conductor connected to the other boom, or replaced by an integrated stripline in a multi-layer antenna configuration, as shown in Fig. 1.7. However, neither of the two approaches would scale up to millimeter-wave region due to the tiny antenna dimension and the high impedance stripline radiation. At millimeter-wave frequencies, the development of LPDA antennas has been mainly limited by fabrication capabilities. In [70], an LPDA antenna is built on a GaAs substrate with a

![Fig. 1.6 Coaxial feed arrangement for an LPDA antenna.](image)
13.4% bandwidth realized around 94GHz. In [71], an LPDA antenna is fabricated using PCB and fed with a substrate integrated waveguide [72]. About 46% bandwidth is obtained at Ka-band. However, due to the fabrication limitation in integrating a feeding line with the antenna, both of the two millimeter LPDA realizations are fed directly from the antenna low frequency end, which significantly limits the antenna bandwidth and adversely affects its far-field.

It is evident that the PCB process is not a good solution for millimeter-wave LPDA antenna development. Even for the easily machinable planar log-periodic antennas, the tiny antenna features are not easily realizable in the standard PCB process. The newly developed surface micromachining technologies, such as E-FAB and PolyStrata [24-25] are potential candidates for antenna fabrication. Due to the rapid growth of the military and satellite applications, millimeter-wave frequency independent antennas are becoming indispensable components in the next generation of electronic attack, radar and satellite sub-systems. It is thus important to design log-periodic antennas with respect to the requirements of the newly developed fabrication processes to achieve consistent and repeatable performance.
1.1.3 Optical Antennas for On-chip Communication

The optical antenna is a relatively new concept that started to gain increased interest at the beginning of this century. They are typically utilized as near-field metallic scatters for near-field enhancement to increase the resolution of scanning near-field optical microscopies [28]. However, these realizations are not suitable for signal transmission and reception as transmission lines are not readily available to connect these metallic antennas. They are also lossy and exhibit strong plasmonic effect which significantly shifts the antenna resonant frequencies. However, efficient optical antennas that can be used as on-chip optical signal transceivers are desired for next generation multi-core microprocessors.

As the density of silicon complementary metal-oxide semiconducting (CMOS) transistors continues to increase at a fast pace, the power consumption of the traditional electrical interconnect fabric between cores may well exceed the projected on-chip communication power budget for the future technology generations [73]. A new interconnection network that allows integrations of tens or even hundreds of cores becomes critical for further development of multi-core microprocessors. Recent developments in silicon photonics enable a promising alternative, specifically a nanophotonic interconnect [74]. Using on-chip optical waveguides, rings, couplers, etc, a variety of interconnect components have already been developed. Substantial improvements over electrical interconnects in throughput, latency, and power efficiency have been demonstrated by simulations based on nanophotonic device research. However, even for nanophotonic interconnect, the latency-critical coordination messages are not handled well [75]. Coordination messages that require broadcast or multicast employ multiple circuits and retransmissions for each coordination transaction, which increases not only signal latency but also power consumption. To address this issue a novel optical interconnect approach based on
on-chip optical antennas is proposed in [75-76]. Dramatic improvement is predicted in cache miss latency and network power consumption compared to both electric and conventional point to point nanophotonic solutions. However, technology compatible and efficient optical antennas for on-chip signal transmissions have yet to be demonstrated.

Due to the various disadvantages of the metallic optical antennas, it is critical to design optical antennas using other materials on wafer environments compatible with other nanophotonic devices used for the core-to-core interconnect. The challenges around antenna shaping, feeding, bandwidth, radiation control, and measurement, need to be addressed imperatively for proper operation for the specific application. The many-channel optical antenna based signal hub for data broadcast and exchange should also be designed, optimized and tested to work with the other nanophotonic components for the next generation many-core processor core-to-core interconnections.

1.2 Dissertation Organization

This thesis is organized as follows:

Chapter 2 discusses a novel mechanically robust antenna feeding method to increase the bandwidth of the microwave patch antennas. Two antenna prototypes are designed, fabricated and tested. After a description of antenna structures, a detailed parametric study is conducted following by the comparisons between the simulated and measured antenna performance. The effects of the antenna structural parameters on the impedance matching, port coupling and gain are studied. The sidelobes and non-ideal hybrid effect are also discussed.

Chapter 3 outlines the development of millimeter-wave log-periodic antennas. Two end-fire LPDA antennas are designed and fabricated using a sequential surface micromachining
technology PolyStrata. Due to the layer thickness uncertainties in different PolyStrata runs, a
generalized antenna model is developed by using nine 100µm uniform strata layers, which can
be easily implemented in real fabrications with respect to the specific layer configurations. The
antenna structural parameters, such as the growth rate, spacing factor, and unbalanced dipole
configurations, are thoroughly studied. A band rejection filter which can be monolithically
integrated within the antenna is also designed for 60GHz rejection. Following that, two antenna
realizations manufactured in different fabrication runs are demonstrated. Measurements up to V-
band are conducted to fully validate antenna performance. Also, a planar millimeter-wave log-
periodic antenna is designed and fabricated using conventional PCB process. Dual-polarized
operation and reduced cost compared to the PolyStrata fabrication are demonstrated. Antenna
measurements performed up to 40GHz are also provided.

Chapter 4 discusses wideband on-chip optical dielectric antennas for core-to-core
communications for the next generation microprocessors. The antenna geometry and the wafer
specifications are given with simulated results based on two different computational methods.
Following that, the measurement setup and measured transmission results between separated
optical antenna pairs are provided. The function of the polymer coating on top of the wafer is
also discussed. The layouts of the two optical signal hubs composed of the demonstrated optical
antennas are developed. The operating principle of the signal hubs is established with
comprehensive measurements as validations. To further verify the hub operating theory, a
microwave wireless interconnection network based on Vivaldi antennas is also designed,
fabricated and measured. Excellent agreement between theory and measurement is obtained.

Finally, Chapter 5 overviews the thesis, summarizes the contributions and outlines
several possible directions for future work.
2.1 Introduction

Patch antennas are a popular choice for various communication systems due to their low cost, low profile and easy for fabrication properties, etc. However, an inherently narrow bandwidth is an issue that inhibits their use for many commercial and defense applications. To address this bottleneck, various single and stacked layer patch configurations have been proposed over the past two decades. Common single layer patch approaches for bandwidth enhancement revolve around radiator shaping and feed design. Modified patch shapes force the currents to flow on different paths thus resulting in multiple resonances. U-slot [48-51] and E-shaped patches [52-53] are appropriate examples. Typically, these patches are fed by a probe directly connected to the patch, and achieve 30%~40% bandwidths when air loaded. The main problems with these approaches include inherently high cross polarization and susceptibility to mechanical vibrations when patch is suspended in the air. Capacitively coupled suspended patch antennas, such as the L-probe fed patch [54-69] are also proposed for bandwidth enhancement. The L-shaped feed introduces a resonance that is offset from the internal TM mode patch resonance. Impedance bandwidths of about 30% are demonstrated. Additional probe shape modifications, such as T-shaped probe [60-61], further increase antenna impedance bandwidth to
about 40%. However, for most of these approaches, the pattern is quite inconsistent and often not discussed.

To improve pattern symmetry and its consistency, a set of two differentially excited probes is preferred [62]. With a single probe feed, the co-polarized radiation pattern undergoes small to moderate squint and a large dipole like cross-polarization occurs in the H-plane at higher frequencies. The increased electrical size of the patch and significantly altered current distribution on non-radiating edges are the main causes of the pattern degradation. Also, when the electrical length of the vertical probe approaches quarter wavelength, the probe starts to radiate as a top loaded monopole. With two differentially fed probes, current components on non-radiating edges as well as the probe radiation are canceled out due to the strict enforcement of the differential phasing. As a result, the cross-polarization level is greatly reduced, and the in-band pattern consistency is well maintained.

In this thesis, two embodiments of the capacitively coupled differentially fed single patch antennas are proposed. Two thick cylindrical probes excite the antenna and firmly hold the patch above the ground plane. This enables the necessary capacitive coupling and excellent mechanical stability. Thick cylinders have smaller inductance compared to the traditional L or T shaped probes, thus enabling larger impedance bandwidths. Specifically, the baseline configuration referred in this chapter as Antenna 1 has VSWR<2:1 from 1.3GHz to 4.1GHz (103%). Broadside gains above 5dBi and 8dBi are obtained over 2.0GHz to 4.1GHz (68%) and 2.9GHz to 3.9GHz (29%), respectively. To reduce coupling losses, the baseline configuration is modified by introducing a metallic plate in the middle of the patch between the probes. This embodiment, referred in this chapter as Antenna 2, has coupling lower than -11dB throughout the investigated bandwidth. The impedance bandwidth with VSWR<2:1 is from 2.3GHz to 4.3GHz (60%).
Broadside gains above 5dBi and 8dBi are obtained over 2.2GHz to 4.1GHz (60%) and 2.4GHz to 3.9GHz (48%), respectively. The overlapping bandwidths for gain above 5dBi and above 8dBi, with VSWR<2:1 and port coupling below -11dB are 56% and 48%, respectively. Both antennas have cross polarization levels below -20dB throughout their 3dB beamwidths and bandwidths. Radiation patterns are stable with 3dB beamwidth variations of ±11° and ±9° in E-plane and H-plane, respectively. A finite element method code Ansoft HFSS [77] is used for the analysis and design. The prototype antennas are fabricated and theory is verified through measurements.

This chapter is organized as follows: Section 2.2 outlines the two proposed antenna configurations. The results for impedance, coupling and radiation patterns are provided in Section 2.3. The effects of the antenna structural parameters on the impedance matching, port coupling and gain are studied in Section 2.4. The antenna sidelobes and non-ideal hybrid effect are discussed in Section 2.5.

2.2 Antenna Configurations

The baseline configuration is shown in Fig. 2.1. This antenna is designed to work at the center frequency $f_0 = 2.75$GHz ($\lambda = 109$mm). Note that this antenna is not developed for a specific system. Rather, its development was motivated by the desire to enhance the bandwidth of both impedance and pattern. That said, the scaling principle can be successfully implemented to tailor its use for a specific wireless need. The antenna is composed of a square patch and ground as well as two cylindrical probes connected to 50Ω SMA connectors by vias. To provide mechanical robustness, a substrate is used between the patch and the probes as well as between the ground plane and the probes. Metallic disks at the bottom side of the patch substrate and top side of the ground plane substrate are etched to allow soldering of cylindrical probes. A square
patch with $P1=P2=40\text{mm} \left(0.37\lambda_0\right)$ on a $T_p=1.575\text{mm}$ thick FR4 substrate ($\varepsilon_r=4.4$) is supported by two copper cylinders, whose height and diameter are $H=9\text{mm} \left(0.083\lambda_0\right)$ and $D_c=9.5\text{mm} \left(0.087\lambda_0\right)$, respectively. A $G=160\text{mm} \left(1.48\lambda_0\right)$ finite size square ground plane with a $T_g=3.175\text{mm}$ thick Rogers RT5880 substrate ($\varepsilon_r=2.2$) is located under the two cylinders. The cylinders, separated by $M=30\text{mm} \left(0.275\lambda_0\right)$ are equidistant with respect to the patch center. They are used to capacitively excite the top patch and are connected to the SMA connectors as shown.
in Fig. 2.1. Robust attachment of the cylinders to the patch and to the ground plane is achieved by two circular copper disks with diameter of \( D_d = 11 \text{mm} \) that are etched on the bottom side of the patch substrate as well as the top side of the ground substrate. The two cylinders are then soldered on these disks, thus obtaining excellent structural rigidity.

Antenna 2 is configured similarly as Antenna 1 except that a vertical metallic wall is placed between the two cylindrical probes in H-plane. This wall shorts the patch and the ground plane as shown in Fig. 2.2 to reduce the port coupling. For improved performance, the patch size is changed to \( P_1 \times P_2 = 40\text{mm} \times 31\text{mm} \). The two cylindrical probes have \( D_c = 12\text{mm} \) diameter and height with \( M = 15\text{mm} \) center to center separation.

![Fig. 2.2 Top, side, and front views of the proposed patch Antenna 2 with denoted structural parameters.](image)
2.3 Antenna Performance

2.3.1 Measurement Setup

To evaluate antenna response, ideal equal amplitude/180° out of phase excitations are needed at the two antenna ports. Wideband hybrids have non-zero amplitude and phase imbalance and when included into the measurements introduce far-field pattern degradation. An ideal excitation is obtained by conducting a calibrated two-channel port measurement of complex fields and then post-processing the data for a desired amplitude and phase set. Once this measurement is conducted, any realistic excitation can be easily accounted for by proper weighting.

The VSWR and coupling measurements are conducted on an Agilent’s Vector Network Analyzer 8719ES, while the far-field measurements are taken in a 13.5m long anechoic chamber at Lockheed Martin Space Astronautics in Denver, CO. Numerical results are based on Ansoft’s HFSS.

2.3.2 Antenna 1

Measured port VSWR versus frequency is shown in Fig. 2.3. As seen, the antenna port impedance is matched very well from 1.3GHz to 4.1GHz, which is a 103% impedance bandwidth for VSWR<2:1. A closer inspection of measured scattering parameters reveals that there is a strong coupling between the two ports. This issue is seldom mentioned in previous articles, but it is present and has a profound impact on antenna performance. Shown in Fig.2.4 are measured and computed couplings between the antenna ports. As seen, at 1.3GHz, the port coupling is as large as -1.8dB, indicating that significant amount of available power is lost, i.e. not radiated. Due to this high port-to-port interaction, the good impedance matching at lower
Fig. 2.3 Measured and simulated Antenna 1 VSWR and broadside realized gain.

Fig. 2.4 Measured and simulated Antenna 1 port coupling.
frequencies becomes misleading. To reduce coupling, a shorting wall between the patch and the ground plane is implemented in Antenna 2, as will be described in section 2.3.3.

The simulated and measured broadside realized gains of Antenna 1 are shown in Fig. 2.3. The antenna has realized gain >5dBi from 2.0GHz to 4.1GHz, and realized gain >8dBi from 2.9GHz to 3.9GHz. These correspond to 68% and 29% bandwidths, respectively. Below 2.0GHz, and above 4.1GHz, the antenna gain drops sharply. Gain reduction at lower frequencies is mainly due to the coupling loss. Note that the effective aperture size is also reduced, which in turn lowers the patch directivity. For the frequencies above 4.1GHz, the broadside gain reduction is primarily due to pattern degradation. This will be further discussed in section 2.4.

Frequency variations of 3dB beamwidths in E and H planes are shown in Fig. 2.5. As seen, in E-plane, the 3dB beamwidth varies between 50° and 72° from 2.0GHz to 4.1GHz,
Fig. 2.6 Radiation patterns of Antenna 1 at 2.1GHz, 3.1GHz, and 4.1GHz. Left: Measured, Right: Simulated.
while in H-plane, the variation is from 70° to 84°. Note that the 3dB beamwidths are not plotted for frequencies below 1.3GHz and above 4.3GHz, due to the pattern deformations.

Measured and simulated antenna radiation patterns at three different frequencies, 2.1GHz, 3.1GHz and 4.1GHz, are shown in Fig. 2.6. Both co- and cross-polarized patterns are provided. As seen, good agreement between the measured and simulated results is obtained for the E and H plane co-polarization. Simulated cross polarization is negligible at low- and mid-band frequencies. Measured cross polarization is 25dB below the co polarization values, indicating good polarization discrimination by the antenna. Note that the cross polarization discrepancy is mostly caused by the range imperfections (tower bounce and multipath).

2.3.3 Antenna 2

Antenna 2 uses a vertical shorting wall between the patch and the ground plane to reduce the coupling between antenna ports. Measured and simulated port VSWRs are shown in Fig. 2.7.

![Fig. 2.7 Measured and simulated Antenna 2 VSWR and broadside realized gain.](image-url)
As seen, VSWR<2:1 is obtained from 2.3GHz to 4.3GHz, i.e. over a 60% bandwidth. The computed and measured port coupling are shown in Fig 2.8. Compared to -1.8 dB maximum value for Antenna 1, obtained port coupling is better than -11dB for all frequencies, verifying the importance of the shorting wall in mitigating the negative effects of port interactions.

The antenna broadside realized gain is shown in Fig. 2.7. As seen, Antenna 2 has gain>5dBi from 2.2GHz to 4.1GHz, i.e. 60% gain bandwidth. From 2.4GHz to 3.9 GHz (48% bandwidth), the antenna gain is above 8dBi.

Measured and simulated 3dB beamwidths of Antenna 2 are shown in Fig. 2.9. In E-plane, the 3dB beamwidth varies between 50°and 72° from 2.2 GHz to 4.1 GHz, while in H-plane, the variation is between 76° and 94°. The 3dB beamwidth is not plotted for frequencies below 1.3GHz and frequencies above 4.3GHz due to the pattern degeneration as discussed in subsection 2.4.
The far-field patterns of Antenna 2 are shown in Fig. 2.10. As seen, the introduced wall does not affect the radiation patterns and they are similar to those of Antenna 1. The cross-polarization is consistently below -22dB.

### 2.4 Parametric Study

The effect of the cylindrical probe height (H) on Antenna 1 is plotted in Fig. 2.11. For computed heights, the shape of VSWR is about the same, however, the mid-band peak value decreases gradually with the increase of H. It is also seen that, as H increases, the antenna VSWR and gain shift to the lower frequencies, due to the enhanced E-field fringing at the radiating edges.
Fig. 2.10 Radiation patterns of Antenna 2 at 2.3GHz, 3.2GHz, and 4.1GHz. Left: Measured, Right: Simulated.
The variations of VSWR and gain of Antenna 1 with the probe thickness (Dc) is shown in Fig. 2.12. Note that the copper disk is mainly used to facilitate soldering, and its diameter Dd changes according to Dd=Dc+1.5mm for this study. As seen, the first VSWR dip is insensitive to the changes of Dc, while the second dip slightly varies in its value but not in position. This is expected since the first resonance is from the patch, while the second is associated with the probe. When Dc changes, the patch dimensions are not affected, but the probe inductance as well as the coupling capacitance varies. For the broadside gain, its bandwidth does not vary with the probe diameter, but its maximum value decreases as Dc increases.

As shown earlier, the wall in Antenna 2 plays an important role in reducing the port-to-port coupling. Its length is swept from 31mm to 71mm and results are shown in Fig. 2.13. As the
Fig. 2.12 VSWR versus probe thickness $D_c$ (Antenna 1).

Fig. 2.13 Port coupling versus wall length (Antenna 2).
wall length increases, the port coupling is reduced. As seen, for wall lengths above 61 mm, the maximum coupling value reduces slightly. For this reason, the value of 61mm is adopted for the Antenna 2 prototype.

Antenna 2 has cylindrical probes close to the isolating wall (1.5mm for M=15mm). To study its manufacturing tolerance, M is swept from 14mm to 16mm. As seen in Fig. 2.14,

![Graph showing VSWR & gain versus probe distance M (Antenna 2).](image)

Fig. 2.14 VSWR & gain versus probe distance M (Antenna 2).

different values of M give little broadside gain variations, while VSWR suffers from moderate shifts, indicating moderate sensitivity to parameter M. This antenna can be easily scaled to operate at Ka and above frequencies [78], where repeatable antenna performance can be guaranteed by, for example, the PolyStrata™ process [25].

### 2.5 Discussion
Radiation pattern stability for both antennas deteriorates at higher frequencies. As seen in Figs. 2.6 and 2.10, the sidelobe level increases, broadside gain is reduced, and cross polarization, albeit small, becomes noticeable. A traditional approach in understanding the cause is to look into surface waves, higher order patch modes, and probe radiation. The two lowest order surface waves for chosen substrate have DC and 20GHz cutoffs [79]. If excited, these waves can reach the edge of the ground plane, scatter from there and contaminate the far-field. However, the lowest order mode should affect all frequencies and the first higher order mode is far above the antenna operating bandwidth. The simulated current distributions of the two patch antennas at 4GHz are shown in Fig. 2.15. As seen, Antenna 2 has small currents on the non-radiating edges. For Antenna 1, although horizontal current (higher order mode) shows up at the patch's non-radiating edges, the 180° phase difference between the currents on top and bottom halves of the patch cancels out their radiation. Thus, the surface current inspection disputes any claims that higher order modes cause the far-field degradation. Finally, the probes radiation possibility is
also disproved by simulation. Specifically, when the probe height $H$ is set to be smaller than 1.5mm, high side lobes still occur.

To understand the high frequency pattern degradation, the array theory [80] is revisited. The two radiating edges of the patch are considered as two slot array elements with a distance $P$ between them. Assuming the ground plane is infinite, the application of image theory leads to four array elements with two elements above and two elements below the ground plane. As frequency increases, the electrical distance between the array elements also increases. Based on the array theory, the broadside gain of this four elements array configuration will gradually decrease. At the same time, the end-fire radiation starts to dominate, as observed in our simulations and measurements. To verify this claim, a theoretical array factor for this array composed of four magnetic hertzian dipoles is computed using the fundamental array theory from classical textbooks [80]:

$$
\cos[\frac{\beta \times (P + 2 \times \Delta P)}{2} \times \sin(\theta) \times \cos(\phi)] \times \cos[\beta \times h \times \cos(\theta)]
\times \sqrt{1 - [\sin(\theta) \times \sin(\phi)]^2}
$$

Eq. 2.1

In Eq. 2.1, $P$ is the patch edge length, $h$ is the distance between the patch and the ground plane, $\beta$ is the phase coefficient, and $\Delta P$ is the extra length due to the fringing, herein equal to 6.98mm. Based on Eq. 2.1, the array far-field pattern is plotted in Fig. 2.16. Since the bottom half is below the ground plane, only the top half patterns are considered. As seen, the two side lobes show up at 3.1GHz, and become very large at 4.1GHz. This result correlates very well with measurements.

Aside from the pattern degradation at high frequencies, non-ideal hybrid affects the antenna overall performance. For testing, a poorly designed 180° hybrid is used, which operates
Fig. 2.16 Antenna radiation patterns based on the array theory at 2.1GHz, 3.1GHz, and 4.1GHz.
from 1GHz to 5GHz with a 1.2dB and 20° maximum amplitude and phase misbalances occurring at 3.7GHz and 4GHz respectively, as shown in Fig. 2.17. When it is connected to Antenna 2, the VSWR measured at the hybrid input port is shown in Fig. 2.18. As seen, the VSWR is reshaped due to the hybrid imperfection and coupling between the two antenna ports, while its bandwidth remains the same, which indicates that the hybrid has only small affect to the antenna input impedance. However, the far-field cross polarization in the H-plane is exacerbated, as seen in Fig. 2.19. The reason for that is the differential currents on antenna non-radiating edges start to radiate due to their imbalance. A commercial available wideband hybrid, such as the JSO-08-471 2GHz to 8GHz 180° hybrid from Pulsar Microwave, provides excellent port balance. When their data is taken into our simulation, the cross-polarization can be improved to -29dB at 4.1GHz.

Fig. 2.17 Measured performance of the 180° hybrid
Fig. 2.18 Measured VSWR with and without the 180° hybrid.

Fig. 2.19 Simulated radiation pattern of Antenna 2 synthesized with measured hybrid misbalance at 4.1GHz.
2.6 Conclusion

This chapter presents two novel microwave wideband patch antennas theoretically and experimentally. A feeding technique referred to as differential cylindrical probes is used to obtain 103% impedance bandwidth and an 68% gain bandwidth for Antenna 1. To reduce the port-to-port coupling, Antenna 2 are modified from Antenna 1 by introducing a metallic wall between the two feeding probes shorting the patch and the ground plane. Impedance and gain bandwidths of 60% are well maintained with port coupling suppressed below -11dB. Developed antennas are mechanically robust. The issues regarding the ports isolation, non-idea hybrid effect and consistency of antenna’s far field pattern are discussed in detail and an approach for mitigation thereof is successfully demonstrated.

Although both antennas have large impedance bandwidths, only Antenna 2 has low coupling losses and thus has better performance. For VSWR<2:1, gain>5dBi and port coupling<-11dB, the overlapping bandwidth of Antenna 2 is from 2.3GHz to 4.1GHz (56%). For gain>8dBi, the overlapping bandwidth is from 2.4GHz to 3.9GHz (48%). Table 2.1 shows the summary of the performance of various reported single layer wideband patch antennas. As seen, for all the listed single port single layer patch antennas, the far-field bandwidth for gain>5dBi and cross-polarization <-10dB is below 20%. Although the patch antenna demonstrated by Rao. P.H. obtains 40% impedance and far-field bandwidth by using differentially T-probe feed. The port-to-port coupling is neglected in their study.

It is important to recognize that for Antenna 2 the introduction of the vertical wall increases the fabrication complexity. This wall must be in a direct contact with the patch and the ground plane in order to fully enforce the cancellation of the vertical E field. Since both the patch
Table 2.1 Summary of the performance of reported single layer wideband patch antennas

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<tr>
<td>Impedance bandwidth for VSWR&lt;2</td>
<td>47%</td>
<td>30.3%</td>
<td>36%</td>
<td>19.5%</td>
<td>40%</td>
<td>60%</td>
</tr>
<tr>
<td>Far-field bandwidth for gain&gt;5dBi and X-pol&lt;-10dB</td>
<td>&lt;16%</td>
<td>0%</td>
<td>&lt;20%</td>
<td>19.5%</td>
<td>40%</td>
<td>60%</td>
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<tr>
<td>Port coupling</td>
<td>Single port</td>
<td>Single port</td>
<td>Single port</td>
<td>Single port</td>
<td>n/a</td>
<td>&lt;-11dB</td>
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<tr>
<td>Overlapping bandwidth</td>
<td>&lt;16%</td>
<td>0%</td>
<td>&lt;20%</td>
<td>19.5%</td>
<td>n/a</td>
<td>56%</td>
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and the ground plane utilize thick substrates, the wall assembly becomes more difficult. Nevertheless, the overall mechanical robustness of Antenna 2 remains very good.

Compared to the frequency independent antennas developed in Chapter 3, patch antennas are not conventional wideband antennas due to their operation modes. However, by carefully designing their feeding methodologies, it is proved that the bandwidth of patch antennas can be significantly improved. In microwave frequencies, wideband patch antennas are readily achievable by using the traditional fabrication techniques, such as the herein used PCB process. For millimeter-wave frequencies, recently developed 3D micromachining technologies can be used for millimeter-wave wideband patch antenna development, such as the one used in developing the Log-Periodic dipole array antennas in Chapter 3.
3.1 Introduction

Log-periodic dipole array (LPDA) antennas [63] have broad bandwidth with moderate gain and consistent radiation patterns. When considering their inherent low cost, it is no surprise that these antennas made a huge impact on both commercial and military applications around the world. Detailed theoretical and experimental studies on LPDA antennas carried out soon after their invention [64] provided a foundation for their further development. LPDA antennas consist of two sets of crisscrossed monopoles whose lengths and separations are related via the growth rate $\tau$. The lowest and highest frequencies of operation are determined by the lengths of the longest and shortest dipoles, respectively. Low frequency (VHF/UHF band) LPDA antennas are typically constructed with metallic tubes that are low cost and easy to fabricate [81]. LPDA antennas built using printed circuit boards [65] reduce their physical size and cost while increasing their operating frequency. Printed LPDA antennas fed by stripline, microstrip and coaxial cable attached to the board have been realized up to Ku-band [65-69]. A printed LPDA antenna fed with the recently introduced substrate integrated waveguide (SIW) operates up to 40GHz [71].

In modern electronic warfare and communication systems, wideband millimeter-wave (MMW) antennas are indispensable. An LPDA antenna is a good candidate due to its consistent
performance and low profile. However, the limiting factor of scaling the LPDA antenna up to MMW range is its feed. LPDA antennas need to be fed from their high-frequency end with either an integrated [63-68] or attached [69] TEM line going along the boom to retain their wideband characteristic. Due to their small size, recent MMW LPDA antennas designed on PCB and GaAs substrates are fed from their low-frequency end [70-71], which limits their impedance bandwidth to 46% and 13.4%, respectively. A recently developed sequential surface micromachining technology, known as PolyStrata [25], permits the use of dielectric straps embedded into metallic layers as support fixtures, thus enabling coaxial line structures to be monolithically integrated with the designed devices [82-84].

In this chapter, two MMW LPDA antennas with monolithically integrated micro-coaxial line feed are demonstrated. The antennas are designed and fabricated using PolyStrata runs with different layer configurations. A micro-coaxial line based feed is designed and monolithically integrated within both antennas to provide band rejection at 60GHz and further emphasize the potential of this technology for MMW uses. A generalized 9-strata LPDA design procedure for photo-lithography based (and enabling) manufacturing is discussed first. The design procedure is then implemented in two different runs to realize the LPDA antennas. Antenna 1 is designed to operate from 18GHz to 110GHz with 60GHz band rejection. VSWR<2.5 is obtained using probe measurement from 10GHz to 40GHz. Antenna 2 is designed to operate from 18GHz to 50GHz with the same feed used to sharpen its high-frequency cutoff. Impedance and far-field measurements for Antenna 2 are conducted throughout its bandwidth. VSWR<2.5, consistent patterns with nominal E and H-plane 3dB beamwidths of 56º and 72º, respectively, and 10dBi directivity are obtained.
This chapter is organized as follows: Section 3.2 briefly outlines the PolyStrata Process. Section 3.3 provides the design procedure and relevant parametric studies of an LPDA antenna using a generalized strata configuration. Section 3.4 discusses the design and performance of the integrated antenna feed for 60GHz rejection. The antenna measurements are described in Section 3.5.

3.2 PolyStrata Fabrication

In PolyStrata, micron size structures are fabricated with a sequence of standard photolithographic steps on a substrate carrier, as depicted in Fig. 3.1. Each copper layer is chemically

![Fig. 3.1 The sequence of fabrication steps applied to build five layer recta-coax lines and related components. (Lithography: develop photo resist to form channels for the copper walls of the structure, Electroplating: grow the copper where the photoresist is not present, Repeat: this process is repeated many times for multiple strata configurations, Release: photoresist is removed chemically, leaving an air filled coax line with periodic dielectric straps.]

polished before the next layer is deposited. Once the structure is built, the resist is drained through the openings in the top and side walls, which are referred to as release holes. The unique feature of this fabrication technology is the capability to embed dielectric straps in certain layers, and use these to support suspended metallic structures, such as the center conductor of a coaxial line. The thickness of each strata layer typically ranges from 20µm to 125µm with 10 or more layers easily machined. Examples of the layer thickness from two fabrication runs (PS4 and PS5) used for the design of the herein proposed antennas are listed in Table 3.1. 20µm thickness of

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*Strata 5 is split into two 50 µm layers with different masks

dielectric straps can be inserted under S5a for PS4, and S3 and S7 for PS5. Note that neither PS4 nor PS5 are tailor made for our needs, instead, they are provided for shared multi-user runs. To provide a general design guideline not only limited to PolyStrata but for other current or future thick photo-lithography micromachining based MMW LPDA antennas, constant 100µm thick layers are used in the design process. This procedure is then implemented for two PolyStrata configurations with slight parameter changes to reduce the added losses introduced from layer thickness variations.
3.3 Generalized LPDA Antenna Model

3.3.1 Boom Line

A drawing of the antenna boom line with nine strata layers is shown in Fig. 3.2. Layers 1-5 are used to form the lower boom, where a coaxial line is integrated. Layers 6-9 form the solid upper boom. Note that the number of layers used for the lower boom must be sufficient to form a closed recta-coaxial line. For mechanical reasons, the two booms are connected at the antenna’s low-frequency end by plating a Bc=3mm long beam in the sixth layer. As will be shown in
Section 3.3.4, this short is far-enough from the antenna’s low-frequency active region, therefore its effect on antenna performance is negligible [64]. To reduce the reflections compared to a rectangular shaft at the feed region, a cylindrical post is used to connect the center conductor of the micro-coaxial line to the antenna’s upper boom. The boom width is tapered from Bo1=0.66mm to Bo2=0.5mm, whereas its length BL=24.75mm is determined by the antenna parameters. The 50Ω input impedance is obtained by carefully choosing the values of Bi1 and C1. The values of Bo2, Bi2 and C2 are specified so that the coaxial line impedance is the same as that of the boom line at the feed region. Note that even though the antenna impedance is determined by the combination of the dipole and boom impedance, it is found that the antenna impedance is very close to the boom impedance when the dipole length is comparable to the boom width, as will be shown in Section 3.3.4. The diameter of the cylindrical post is the same as the width of the boom at the feed region. The distance between the boom and the cylindrical post center, D, is fine-tuned to improve return loss.

3.3.2 LPDA Antenna

A drawing of the LPDA antenna is shown in Fig. 3.3. The antenna growth rate \( \tau \) and spacing factor \( \sigma \) are specified to be 0.92 and 0.1, respectively, to compromise the antenna length and far-field performance as will be discussed in Section 3.3.4. The number of the dipole elements \( N=32 \) is determined by [80]

\[
N = 1 + \frac{\ln(B_s)}{\ln(1/\tau)} \quad (3.1)
\]

where \( B_s \) is the design bandwidth, which is given by

\[
B_s = \frac{f_{\text{max}}}{f_{\text{min}}} [1.1 + 7.7(l - \tau)^2 \cot \alpha] \quad (3.2)
\]
Fig. 3.3 Drawing of LPDA antenna design ($L_{\text{max}}=8$ mm, $\tau=0.92$ and $\sigma=0.1$) with magnified low and high-frequency regions.

$f_{\text{min}}$ and $f_{\text{max}}$ are the operating frequencies for the largest and smallest dipoles, which are specified to be 18GHz and 180GHz, respectively. $\alpha$ is the apex half angle, defined by

$$\alpha = \tan^{-1}\left[\frac{1-\tau}{4\sigma}\right]$$ (3.3)

Note that the desired antenna operating band is 18GHz-110GHz. However $f_{\text{max}}=180$GHz is chosen to improve the antenna matching as will be discussed in Section 3.3.4. It is expected that the antenna will not have consistent patterns above W-band due to the large boom dimension. The length of the longest dipole determined by the 18 GHz cutoff its fixed at $L_{\text{max}}=8$mm (0.48$\lambda$ at 18GHz). The length of the shortest dipole is calculated based on $L_{\text{max}}$, $\tau$ and N to be
$L_{\text{min}} = 0.603\,\text{mm}$, while the width of each dipole is kept at 1/8 of its length to satisfy the aspect ratio at high-frequency end. The distance between the shortest and the longest dipole is determined by

$$L = \frac{L_{\text{max}}}{2} \left( 1 - \frac{1}{B} \right) \cot \alpha$$

which is 18.5 mm. The largest dipole and the boom connection area are separated by $B_d = 3.25\,\text{mm}$.

### 3.3.3 Full-Wave Simulation

Time domain finite integration technique (FIT) code CST Microwave Studio [85] and frequency domain finite element method (FEM) code Ansys HFSS [77] are employed for full wave simulations. The simulated antenna VSWR, realized gain in the end-fire direction and

![Fig. 3.4 Simulated antenna VSWR, realized gain and front-to-back ratio.](image)

FIG. 3.4 Simulated antenna VSWR, realized gain and front-to-back ratio.
front-to-back (FB) ratio are shown in Fig. 3.4. The antenna responses from 110GHz to 180GHz are also shown for reference. As seen, the generalized antenna model has VSWR<1.5 and end-fire gain > 7.5dBi within the desired 18GHz to 110GHz band (with maintained consistency up to 180GHz). The antenna FB ratio is > 20dB throughout the bandwidth and drops below 20dB at 124GHz. The simulated antenna radiation patterns at 20GHz, 60GHz, 100GHz, 140GHz and 180GHz are shown in Fig. 3.5. Both the antenna side lobes and cross polarization increase with frequency. The pattern degradation at high frequencies is due to the electrical size of the boom and the connecting cylindrical post height (0.2λ at 124GHz). Even though \( f_{\text{max}} = 180\text{GHz} \) is used, the antenna far-field operational band is limited up to W-band and no claims of wider bandwidth are made herein. Note that the excellent agreement between two fundamentally different full wave methods validates the performance of the designed antenna.

### 3.3.4 Parametric study

The antenna response for growth rate \( \tau = 0.9, 0.92 \) and 0.94 are shown in Fig. 3.6. Note that for different \( \tau \), the spacing factor \( \sigma \) is kept the same with all the other parameters determined by equations (3.1~3.4). As seen, different \( \tau \) values have no pronounced effect on antenna VSWR, but larger \( \tau \) gives higher gain and better FB ratio. \( \tau = 0.94 \) achieves 1~2dB higher gain than \( \tau = 0.9 \) throughout the bandwidth with significant FB ratio improvement below 80GHz. However, the number of elements \( N \) for \( \tau = 0.9 \) is 26, while that for \( \tau = 0.94 \) is 42, indicating the boom length for \( \tau = 0.9 \) (21.25mm) is shorter than that for \( \tau = 0.94 \) (30.75mm) for about 30%.

The antenna responses for spacing factor \( \sigma = 0.075, 0.1 \) and 0.125 with \( \tau = 0.92 \) are shown in Fig. 3.7. As seen, the antenna’s VSWR is close for different \( \sigma \) values. The realized gain has 0.5~2dB improvement for \( \sigma = 0.125 \) compared with \( \sigma = 0.075 \) throughout the bandwidth. For
Fig. 3.5 Simulated antenna radiation patterns at 20GHz, 60GHz, 100GHz, 140GHz and 180GHz.
Fig. 3.6 Simulated (FIT) antenna VSWR, gain and front-to-back ratio for $\tau=0.9$, 0.92 and 0.94.

Fig. 3.7 Simulated (FIT) antenna VSWR, gain and front-to-back ratio for $\sigma=0.075$, 0.1 and 0.125.
front-to-back ratios below 60GHz, $\sigma=0.075$ has a more consistent response than $\sigma=0.125$, while above 60GHz $\sigma=0.125$ delivers 2~5dB better performance. The number of elements N is fixed at 32 for all three cases, since it is not sensitive to $\sigma$ variations based on equations (1-3). The boom lengths for $\sigma=0.075$ and 0.125 are 20.25mm and 29.25mm, respectively.

The VSWR and realized gain of LPDA antennas designed for $f_{\text{max}}=180$GHz ($L_{\text{min}}=0.603$mm), $f_{\text{max}}=90$GHz ($L_{\text{min}}=1.175$mm) and $f_{\text{max}}=45$GHz ($L_{\text{min}}=2.290$mm) are shown in Fig. 3.8. Note that the number of elements N and boom length BL are decreased according to equations (1-4) (N=24 and 16 while BL=23.25mm and 20.25mm for $f_{\text{max}}=90$GHz and 45GHz, respectively). The FB ratio comparison is shown in Fig. 3.9. As seen, the antenna has better VSWR and FB ratio for larger $f_{\text{max}}$. For $f_{\text{max}}=45$GHz, the antenna is only matched up to 43GHz.

![Fig. 3.8 Simulated (FIT) antenna VSWR and realized gain for $f_{\text{max}}=45$GHz, 90GHz and 180GHz.](image-url)
For $f_{\text{max}}=90\text{GHz}$ and $180\text{GHz}$, the antennas have good matching within their bandwidths and gain $> 7.5\text{dBi}$. The reason for the degraded antenna performance for smaller $f_{\text{max}}$ is that the integrated coaxial line impedance is designed to match the boom line impedance. For larger $f_{\text{max}}$, $L_{\text{min}}$ is close to the boom width at the high-frequency end ($L_{\text{min}}/\text{Bo}_2=2.35$ and $1.21$ for $f_{\text{max}}=90\text{GHz}$ and $180\text{GHz}$, respectively), and the antenna impedance is dominated by the boom impedance for these two cases. For $f_{\text{max}}=45\text{GHz}$ ($L_{\text{min}}/\text{Bo}_2=4.6$), the dipoles start to influence the antenna impedance in a greater way. In this case, Bo2, Bi2 and C2 can be tuned to match the coaxial line impedance to the antenna impedance [80], however, this dimension variation is restricted by the permissible fabrication aspect ratio. Note that the boom length for $f_{\text{max}}=180\text{GHz}$ is only $1.5\text{mm}$ longer than that for $f_{\text{max}}=90\text{GHz}$, however, all three parameters studied have better performance for $f_{\text{max}}=180\text{GHz}$ throughout the bandwidth. When a specific layer configuration is implemented, higher than required $f_{\text{max}}$ should be used to improve the antenna performance.

Fig. 3.9 Simulated (FIT) antenna front-to-back ratio for $f_{\text{max}}=45\text{GHz}$, $90\text{GHz}$ and $180\text{GHz}$. 
The antenna VSWR at the cutoff region for $B_d=0\text{mm}$, $3.25\text{mm}$ and $\infty$ (shorting beam removed) are shown in Fig. 3.10. As seen, with the connecting beam between booms at the antenna low frequency region, VSWR at cutoff shifts away slightly from that without the beam. Note that the shorting beam at the antenna lowest frequency region works as a reactive load. For different $B_d$, this load exhibits different impedance to the antenna largest dipole. However, since the current and voltage decay quickly away from the lowest active region, this load shows insignificant impact to the antenna input impedance, especially for the frequencies above 20GHz.

![Fig. 3.10 Simulated (FIT) antenna VSWR for $B_d=0\text{mm}$, $3.25\text{mm}$ and $\infty$.](image)

### 3.3.5 Discussion

Due to the restricted dielectric strap location and the layer thickness variations in different PolyStrata runs, the thickness of the monopoles on the lower and upper boom may not be the same, leading to an unbalanced dipole array configuration. To determine the impact, the
thickness of the 5th layer which contains one half of the monopoles is reduced to 50µm, as shown in Fig. 3.11. The simulated VSWR and realized gain in end-fire direction are shown in Fig. 3.12.

Fig. 3.11 Drawing of the high-frequency region of the generalized antenna model with the 5th layer thickness reduced to 50 µm.

Fig. 3.12 Simulated (FIT) antenna VSWR and realized gain for balanced and unbalanced dipole arrays.
As seen, neither parameter is detrimentally affected by the unbalanced dipole configuration. The antenna beam squint and 3dB beamwidth comparisons between the balanced and unbalanced cases are shown in Fig. 3.13. The unbalanced dipole array configuration introduces up to 4° additional beam squint in both E and H planes, while the 3dB beamwidths in both planes are well maintained.

The antenna gain roll off from 18GHz to 15GHz is < 10dB as shown in Fig. 3.4. One approach to increase the gain roll off is to use a large antenna growth rate $\tau$ and connect the booms right before the longest dipole (Bd=0mm) to choke the active region at the lowest frequency. However, $\tau$ is also related to the boom length, thus a larger $\tau$ results in a longer antenna. For $\tau=0.96$, the boom length is 43.25mm, which is almost twice the boom length when $\tau=0.92$. To shorten the antenna while maintain sharp gain roll off at cutoff, a non-uniform $\tau$ as with quasi-log-periodic designs [81] can be used. In such case, the number of dipole elements N
is determined by $\tau_{\text{avg}}$, the geometric mean of $\tau_{\text{max}}$ and $\tau_{\text{min}}$. Due to the smaller $\tau$ at one side of the antenna, the antenna length is significantly reduced. For example, for $\tau$ varying from 0.96 to 0.89 from the lowest to the highest frequency ends, the length of the antenna is only 28.75mm. Fig. 3.14 shows VSWR and realized gain comparisons between an LPDA antenna with $\tau=0.96$~0.89 and $\tau=0.92$ and $Bd=3.25$mm.

![Simulated (FIT) antenna VSWR and realized gain comparison](image)

Fig. 3.14 Simulated (FIT) antenna VSWR and realized gain for $\tau=0.96$~0.89 and $Bd=0$mm and $\tau=0.92$ and $Bd=3.25$mm.

and $Bd=0$mm, and an antenna with constant $\tau=0.92$ and $Bd=3.25$mm. As seen, within the desired band > 20dB gain roll off is achieved from 18GHz to 15GHz without deteriorating the VSWR and gain.

### 3.4 LPDA Antenna Feed Design

The 60GHz band is generally used for short distance communications due to high signal absorption from oxygen molecules in the atmosphere[86]. To create a band rejection region
around 60GHz for the proposed LPDA antenna, the straight 50Ω recta-coaxial line feed in the antenna's low boom is redesigned based on the filter topology developed in [87]. It is important to recognize that the integration of a wideband feed without band rejection is straightforward and easy. Herein the proposed feed is unique to PolyStrata and can enable additional functionality to the antenna not readily achievable with other technologies in this or lower frequency bands. The circuit model and drawings of the feed are shown in Fig. 3.15 and 3.16, respectively. Six series

![Circuit Model Diagram]

Fig. 3.15 Schematic layout of the antenna feed. ($Z_1=45.6\ \Omega, Z_2=33.9\ \Omega, Z_B=30.9\ \Omega, Z_R=50.6\ \Omega$ and $\lambda=9.7\text{mm}$)

![3D View and Cut Plane Diagram]

Fig. 3.16 3D view (top) and cut plane (bottom) of the antenna feed designed based on Polystrata. The launcher on each end is designed for probe impedance measurements.
and two shunt quarter wavelength sections are used. The length and width of the outer conductor of each quarter wavelength section is 2.5mm and 0.7mm respectively, which is small enough to be integrated in the antenna as shown in Fig. 3.17 (Antenna 1 built in PS4). The width of each center conductor is determined by the corresponding line impedance. Excellent agreement between FIT and FEM simulated results is obtained, as shown in Fig. 3.18. The isolated antenna feed has $|S_{11}| < -10$dB between 12GHz to 50GHz and 73GHz to 112GHz bands.

### 3.5 LPDA Antenna Measurements.

The LPDA model discussed in Section 3.3 is modified based on PS4 and PS5 strata configurations shown in Table 3.1. Fig. 3.19 shows the final structures of recta-coax feed transitions of both antennas. The location of the coaxial line’s center conductor is restricted to layer 5 in PS4 and layer 3 in PS5 because of the dielectric strap locations predetermined by the foundry. The desired operational bands for Antenna 1 and Antenna 2 (feed is excluded) are 18GHz to 110GHz and 18GHz to 50GHz, respectively. To achieve more consistent impedance and patterns over the desired bands, based on the findings from Section III, $f_{\text{max}}=180$GHz and 75GHz are selected for Antenna 1 and 2, respectively. The diameter of the cylindrical post and
Fig. 3.18 Simulated S-parameters of the antenna feed fabricated in PS4.

Fig. 3.19 High-frequency regions of Antenna 1 (top) and Antenna 2 (bottom) built in 11 strata PS4 and PS5 multi-user runs.
boom to post center distance $D$ are reduced to 0.4mm and 0.3mm, respectively, to minimize the added mismatch introduced by the changed layer configuration. The inner conductor width $C_2$ for Antenna 2 is increased to 0.12mm for the same purpose. The growth rate $\tau$ for both antennas varies from 0.96 to 0.89 from the longest to the shortest dipole. Fig. 3.20 shows the low-frequency region of the two antennas. As seen, both antennas have the designed feed integrated for 60GHz band rejections. A probe launcher is designed for Antenna 1 for impedance measurement. A CB-CPW to micro-coaxial line transition is designed and integrated with Antenna 2. Each antenna is attached to a 30mil Rogers RO4350 carrier-board to ease impedance (Antenna 1 and 2) and far-field (Antenna 2) measurements. The CB-CPW to micro-coaxial line transition of Antenna 2 is connected to the CB-CPW line on the RO4350 board. A Southwest Microwave 2.4mm connector is attached to the board upon the carrier assembly to enable Antenna 2 measurements. The fabricated antennas are shown in Fig. 3.21.

![Fig. 3.20 The low-frequency regions of Antenna 1 (top) and Antenna 2 (bottom) built in 11 strata PS4 and PS5 multi-user runs.](image-url)
Fig. 3.21 Photograph of Antenna 1 (top) and Antenna 2 (bottom) fabricated in PS4 and PS5 placed on a carrier-board.

Antenna 1 is measured using a 250µm pitch ground signal ground probe. Due to the upper frequency limit of the available network analyzer at NIST, Antenna 1 is measured only to 40GHz. Obtained result is shown in Fig. 3.22. The fabricated antenna has VSWR<2.5, slightly higher than simulated and likely caused by small layer thickness changes (10% tolerance) [88]. Also noticeable is that the presence of the carrier-board leads to some VSWR degradation around the cutoff frequency. The simulated results indicate strong rejection at 60GHz while good correlation thereof with measurement (up to 40GHz) provides necessary validation.

Measured and simulated VSWR of Antenna 2 are shown in Fig. 3.23. Good agreement is obtained through 50GHz. Similar to Antenna 1, the carrier-board only slightly affects the antenna matching around cutoff. The 60GHz band (shown in inset) is also rejected by the
Fig. 3.22 Measured and simulated (FIT) VSWR of Antenna 1.

Fig. 3.23 Measured and simulated (FIT) VSWR of Antenna 2.
Fig. 3.24 Measured and simulated (FIT) directivity and realized gain of Antenna 2.

integrated feed and the 50GHz cutoff is sharpened, as seen from the simulated results. Fig. 3.24 shows the measured and simulated directivity and simulated realized gain. Good agreement is observed. The directivity stays around 10dBi throughout the bandwidth. Directivity and pattern measurements were performed in an anechoic chamber capable of measurements up to 110GHz. 3-D patterns were taken up to 50GHz (limit of 2.4mm connector) with the use of an automated spherical scanning system. An aluminum measurement fixture is used as the antenna mount to the positioner, as shown in Fig. 3.25. Absorber material (not shown in the figure) is placed between the aluminum plate and the antenna to reduce the back lobe. The addition of absorber reduces the overall radiated power (for the amount in the back lobe), thus slightly increasing the measured directivity. Also shown in Fig. 3.24 are the simulated realized gains with and without the carrier-board. The losses from the mismatch and copper material lower the
Fig. 3.25 Antenna 2 mounted on a metallic measurement fixture. For far-field measurements, the absorber is added directly to the plate.

antenna gain by about 1dB. When the 30mils RO4350 board is attached, the antenna gain has additional 1dB gain drop. The measured and simulated 3dB beamwidths of Antenna 2 are shown in Fig. 3.26 and good agreement is obtained. The antenna has consistent E and H-plane 3dB beamwidths around 56º and 72º, respectively. Measured and simulated antenna radiation patterns at 20GHz, 28GHz, 36GHz, and 45GHz are shown in Fig. 3.27. Good correlation between simulation and measurement is obtained. The measured radiation intensity is reduced from θ = 90º to 270º due to the back-side absorber. The end-fire direction cross-polarization level is better than 15dB throughout the bandwidth.

3.6 Millimeter-Wave Planar Log-Periodic Antenna

The so far discussed LPDA antennas are inherently linearly polarized. For circular polarization two LPDA antennas have to be orthogonally aligned in the center [89] in a
configuration not achievable using conventional surface micromachining technologies. However, multi-arm planar log-periodic antennas can easily operate as linearly or circularly polarized antennas by exciting the antenna arms with different phase combinations. Microwave multi-arm planar log-periodic antennas can be easily fabricated using PCB processes for dual-linear and dual-circular polarizations [90], however, their MMW extensions are more challenging and have yet to be studied.

The PolyStrata surface micromachining process used for the LPDA antenna fabrications is shown to be a reliable manufacturing approach for MMW components. However, its high cost, longer fabrication time and restricted aspect ratio are major drawbacks compared to the traditional PCB process. Considering that nowadays PCB processes can readily achieve 3mils wide traces and trenches [91] with 4mils minimal thickness of low-loss dielectric materials such as RO4350 [92], it is worth exploring their suitability for the design of wideband MMW
Fig. 3.27 Measured and simulated (FIT) radiation patterns of Antenna 2 at 20GHz, 28GHz, 36GHz and 45GHz.
antennas. This sub-section demonstrates that for planar MMW components, such as planar log-periodic antennas, PCB process is a low-cost alternative to surface micromachining.

### 3.6.1 Antenna Configuration and Full-Wave Analysis

In this sub-section, an 18GHz to 40GHz 4-arm planar log-periodic antenna is designed and fabricated using standard PCB process. A drawing of the designed antenna is shown in Fig. 3.28. Two stacked 8mils Rogers RO4003 boards (three copper layers) are used to build the antenna. The D=12mm diameter four-arm planar LP structure lies in the middle layer, with two microstrip lines located on the top and bottom layers, respectively, feeding the antenna through plated via holes. Due to the high nominal impedance of a 4-arm self-complementary log-periodic antenna [93], a three section impedance transformer is integrated with both feeding lines and non self-complementary configuration is used ($\beta_1=25^\circ$, $\beta_2=10^\circ$) [81] to provide better antenna matching to 50\,$\Omega$. Each of the antenna arms features 22 crisscrossed monopoles with a growth rate of 0.91 and 1:1 metal to slot ratio for optimal far-field performance within the above mentioned fabrication limitations.

The simulated antenna S-parameters are shown in Fig. 3.29. As seen, within the 18GHz to 40GHz bandwidth, S11 and S22 < -7dB (VSWR<3) and S21 and S12 < -15dB are obtained. Even though 160\,$\Omega$ to 50\,$\Omega$ impedance transformers are used, the high antenna impedance (288\,$\Omega$ [93]) still renders relatively high reflection coefficients with respect to a 50\,$\Omega$ impedance. The two antenna ports can either be excited separately to achieve two orthogonal linear polarizations, or fed using a quadrature hybrid to obtain two opposite circular polarizations. For both cases, antenna boresight gains (theta=0\,$^\circ$) around 5dBi and consistent 3dB beamwidths between 60\,$^\circ$ and 70\,$^\circ$ are achieved throughout the bandwidth, as shown in Fig. 3.30 and Fig. 3.31, respectively. The simulated antenna radiation patterns are shown in Fig. 3.32. Note that the linearly polarized
Fig. 3.28 Geometry (top) and drawing of the feeding region (bottom) of the designed planar log-periodic antenna.
Fig. 3.29 Simulated (FIT) antenna S-parameters.

Radiation patterns are obtained with port 1 excited and port 2 loaded with 50Ω impedance. Due to the structural symmetry, the radiation patterns from port 2 are similar and not shown here. As seen, good quality bidirectional radiation patterns are obtained with low cross polarizations in both E- and H-plane.

3.6.2 Antenna Measurement

A photograph of the fabricated antenna with attached connectors is shown in Fig. 3.33. The top and bottom boards of the antenna are fabricated separately on the same board. They are aligned and bonded using the set of dowel pins and nylon screws, respectively. The antenna inputs are connected to two Southwest Microwave 2.4mm connectors for measurements. In the far-field measurement setup, an aluminum fixture similar to that shown in Fig. 3.25 is used as the antenna mount to the positioner in the anechoic chamber. Absorbing material is placed between
Fig. 3.30 Simulated (FIT) antenna gain for two linear polarizations (top) and two circular polarizations (bottom).
Fig. 3.31 Simulated (FIT) antenna 3dB beamwidths for two linear polarizations (top) and two circular polarizations (middle and bottom). Note that the antenna is bidirectional and the main lobe in upper hemisphere is used for these plots.
Fig. 3.32 Simulated (FIT) antenna radiation patterns at 18GHz (left), 28GHz (middle) and 40GHz (right). Note that the radiation patterns are obtained with port 1 excited and port 2 loaded with 50Ω impedance.

Fig. 3.33 Photograph of the fabricated antenna with two 2.4mm connectors from Southwest microwave.
the antenna and the aluminum plate to reduce their interaction. To better estimate the real conditions in the antenna measurements, detailed models for two connectors, four nylons screws and the absorbing material are added in the computational model, as shown in Fig. 3.34.

Fig. 3.34 Model of the fabricated antenna used in FIT simulation.

Fig. 3.35 Measured and simulated (FIT) antenna S-parameters.
The measured antenna S-parameters are shown in Fig. 3.35. $S_{11}$ and $S_{22} < -6$dB and $S_{21}$ and $S_{12} < -15$dB are achieved throughout the bandwidth. Since the log-periodic structure is fabricated on the board that contains the 2nd microstrip feeding line and the 1st line is attached to the antenna afterward using screws, the obtained return losses of the two ports are somewhat different. The measured antenna boresight gain, E- and H-plane 3dB beamwidths are shown in Fig. 3.36 and Fig. 3.37, respectively. Since the model used to emulate the real measurements is simplified, the simulated antenna gains are about 1dB higher than measured. Also some moderate ripples in the 3dB beamwidths are not captured in the model. As seen, the two connectors, as well as the nylon screws, impact the antenna 3dB beamwidth consistency significantly. Compared with the ideal case shown in Fig. 3.31, where none of the above scatterers is included, the measured antenna beamwidths vary substantially. It is thus reasonable to expect that a better attachment with integrated feeders will produce better 3dB beamwidth consistency. The measured antenna radiation patterns for port 1 and port 2 are shown in Fig. 3.38 and Fig. 3.39, respectively. As seen, the antenna backside radiation is considerably reduced by the absorbing material used in the measurement. However, due to the discussed scatterers involved in the measurement, when compared to Fig. 3.32, the obtained results show slight degradation across the entire bandwidth. The measurement of the antenna circular polarization performance is not performed due to lack of a wideband quadrature hybrid operating in this frequency band. However, from the above discussed results, similar quality and circularly polarized performance is expected when the antenna is fed by a quadrature hybrid and operates in circularly polarized mode.

Compared to the MMW LPDA antenna, this proposed planar log-periodic antenna has inherent polarization diversity with significantly lowered fabrication cost. However, the antenna
Fig. 3.36 Measured and simulated (FIT) antenna gains.

Fig. 3.37 Measured and simulated (FIT) antenna E- and H-plane 3dB beamwidths.
is bidirectional and in reality a cavity filled with absorbing material should be used to reduce the radiation on the back direction, which reduces the antenna radiation efficiency by half.

3.7 Conclusion

Two wideband micromachined millimeter-wave LPDA antennas are demonstrated in this chapter. An antenna model is developed first to provide a general design procedure for surface micromachined LPDA antennas. This procedure is then implemented for two different PolyStrata configurations to fabricate the proposed LPDA antennas. An antenna feed is designed and monolithically integrated with antennas to give a band rejection at 60GHz. Antenna 1 operates
Fig. 3.39 Measured and simulated (FIT) radiation patterns at 18GHz, 28GHz and 40GHz, when port 2 is excited.

from 18GHz to 110GHz with 60GHz rejection. Measured VSWR<2.5 up to 40GHz and good correlation with simulation validate Antenna 1 operation. Antenna 2 operates from 18GHz to 50GHz with the high-frequency cutoff sharpened by the feed. VSWR<2.5, consistent patterns, E and H plane 3dB beamwidth, and 10dBi directivity are measured throughout the bandwidth. For dual-linear, dual-circular polarizations and lower fabrication cost, a planar log-periodic antenna is designed and fabricated using PCB process. VSWR<3 with boresight gain around 4dBi are obtained in the measurements throughout the 18GHz to 40GHz bandwidth.

Compared to the wideband patch antennas developed in Chapter 2, log-periodic antennas
can achieve much wider bandwidth as expected due to their frequency independent nature. This chapter has demonstrated that the multi-octave wideband LPDA antennas can now be designed for high quality operation deep into MMW region. Additionally, the scalability of the utilized micro-coaxial technology and its ability to monolithically integrate different devices will allow for easy scaling of developed LPDA antennas into sub-MMW band. This work opens research venues that only a few years ago deemed unimaginable.
CHAPTER 4
OPTICAL ANTENNAS FOR ON-CHIP COMMUNICATION

4.1 Introduction

Continuous advances in fabrication technologies have contributed to the reduced size, improved performance, and enhanced power efficiency of transistors. However, metallic wires, the main interconnect solution in integrated circuits (ICs), have not scaled at the same pace. As a result, on-chip communication has become a key challenge for emerging multi-core and multi-billion-transistor ICs. The fundamental physical limitations of electrical interconnects include communication throughput, latency, and power dissipation. For instance, the power consumption of electrical interconnections may exceed the projected on-chip IC communication power budget by over a factor of ten for future technology generations [73].

Recent advances in silicon photonics have enabled promising alternatives for future on-chip interconnections [74]. Based on optical silicon waveguides, various on-chip photonic network designs including bus, ring, mesh and clos network, have been proposed to facilitate efficient on-chip communication for future many-core ICs. However, there are challenges that these optical interconnects have yet to overcome. Particularly, in many-core systems, a vital part of the on-chip traffic are short, often-multicast, and latency-critical messages. They are employed to synchronize concurrent program threads execution, maintain distributed on-chip data coherence, and manage global resources. If these messages are not properly handled, the
many-core IC system performance degrades substantially [94]. Therefore, a power efficient, low latency and high bandwidth optical broadcast communication solution is crucial for providing instant global reach for the performance-critical, multi-destination on-chip traffic.

Foreseeing these challenges, different photonic broadcast communication topologies have been proposed. For example, an N×M Multi-Mode Interference (MMI) network [95] based on the self-imaging theory and equal splitting of the input signal into multiple outputs has been successfully demonstrated. While characterized by small loss and excellent output balance, the MMI network’s bandwidth is inversely proportional to the count of input and output ports [96]. It is shown that 1×32 MMI exhibits a 7nm bandwidth for 1550nm operating wavelength [97]. This inherent property strongly restricts MMI for on-chip broadcast applications where broadband wavelength-division multiplexing (WDM) support is needed. A novel optical interconnect approach based on on-chip wideband optical antennas is proposed in [76]. Dramatic improvement is predicted in cache miss latency and network power consumption compared to both electric and conventional point to point nanophotonic solutions. However, technology compatible and efficient optical antennas for on-chip signal transmissions have yet to be demonstrated. Even though microwave monolithic integrated antennas for intra-chip interconnects have been successfully developed [98-101], reported terahertz and optical on-chip antennas suffer from material loss and require unconventional feed approaches [102-104] to operate as effective on-chip transmitters or receivers.

This chapter demonstrates wideband optical antennas and wideband antenna based optical signal hubs applicable for on-chip interconnect and broadcast. In this chapter, 172 THz (1750 nm) to 222 THz (1350 nm) optical dielectric rod antennas for on-chip interconnect are demonstrated first. They are fabricated on a 200 mm Silicon-On-Insulator (SOI) platform based
on IMEC 193 nm deep UV lithography [105]. A 500 nm thick polymer layer is deposited on the top of the wafer during post processing to function as an asymmetric slab waveguide, thus confining the antenna radiated field to the layer. The antenna is fed directly from an optical waveguide [106] that receives the input signal from a single mode fiber (SMF) through a predefined grating coupler on the wafer [107-108]. Within its 50THz bandwidth, the antenna achieves return loss and end-fire gain greater than 25 dB and 9 dBi, respectively. For the measurements, the antennas are arranged in pairs facing each other with different separations. The transmission measurements are performed from 190 THz to 200 THz and are limited by the bandwidth of the spectrum analyzer and grating couplers. Good agreements with full wave modeling are obtained. Following that, two optical signal hubs with 16 and 32 optical dielectric rod antennas are demonstrated. The hubs operate as data distribution centers, broadcasting signals from an arbitrary antenna to all the others with minimum transmission imbalance, thus allowing low latency wireless links between WDM channels. To construct the hub, antennas are compactly and symmetrically arranged in a circular ring with carefully designed separations. The proposed optical signal hubs operate over a 172 THz to 222 THz bandwidth, which fully satisfies the WDM requirements [75]. The test set-up developed in 190 THz to 200 THz band shows good agreement with numerical modeling. The fabricated 32-channel hub achieves -22.4 dB transmission minimum. This requires a minimum input power of 37 mW to establish the communication between the cores [Appendix C], significantly more efficient than the electrical interconnect alternatives.

The chapter is organized as follows: Sections 4.2.1 and 4.2.2 outlines the proposed optical dielectric rod antenna geometry and its full-wave analysis. Sections 4.2.3 and 4.2.4 discusses the measurement setup and the obtained results. Discussion pertaining to the function
of the polymer coating is given in Section 4.2.5. Sections 4.3.1 ~ 4.3.3 provides the layout, operating principle and performance of two optical signal hubs. Parametric study and the scaled microwave hub measurements are given in Sections 4.3.4 and 4.3.5.

4.2 On-chip Optical Dielectric Rod Antenna

4.2.1 Antenna Geometry

The configuration of the proposed optical dielectric rod antenna is shown in Fig. 4.1. The

![Fig. 4.1 Proposed optical dielectric rod antenna configuration. Top: top and side views. Bottom: 3D model (L1=4 µm, L2=0.5 µm, L3=3 µm, W1=0.45 µm, W2=0.75 µm, W3= 0.13 µm, T1=0.22 µm, T2=2 µm, T3=0.5 µm).]
antenna is designed for intended fabrication on a 200 mm SOI wafer following IMEC 193 nm deep UV lithography guidelines [109]. The SOI wafer features three layers, a \( T1 = 0.22 \pm 0.02 \) \( \mu \text{m} \) top silicon layer (\( \varepsilon_r = 11.9 \)), a \( T2 = 2 \pm 0.05 \) \( \mu \text{m} \) thick Buried Oxide (BOX) layer (\( \varepsilon_r = 2.1 \)), and a 700 \( \mu \text{m} \) bulk silicon substrate (\( \varepsilon_r = 11.9 \)). Note that all the designed structures are developed on the top silicon layer. The antenna is extruded from a \( L2 = 0.5 \) \( \mu \text{m} \) by \( W2 = 0.75 \) \( \mu \text{m} \) by \( T1 = 0.22 \) \( \mu \text{m} \) launcher and exponentially tapered down to a \( W3 = 0.13 \) \( \mu \text{m} \) tip over a \( L3 = 3 \) \( \mu \text{m} \) length. A standard \( W1 = 0.45 \) \( \mu \text{m} \) by \( T1 = 0.22 \) \( \mu \text{m} \) optical waveguide [106] is linearly tapered up over an \( L1 = 4 \) \( \mu \text{m} \) axial length and connected to the launcher. A \( T3 = 500 \) nm polymer layer (\( \varepsilon_r = 2.89 \)) is post developed on the top of the SOI wafer to confine the radiation, as depicted by the grey area in Fig. 4.1 - side view. For the purpose of modeling, the SOI wafer is modeled as a 12 \( \mu \text{m} \) by 16 \( \mu \text{m} \) rectangular cell with the antenna placed in the middle. The excitation is provided by a waveguide fed from the rectangular port at the cell boundary, as shown in Fig. 4.1 -3D model. The bottom bulk silicon substrate is modeled as a 0.5 \( \mu \text{m} \) thick silicon layer, face terminated in a radiation boundary. This reduces the computational complexity while ensures proper substrate functionality.

4.2.2 Antenna Full-Wave Analysis

The time domain finite integration technique (FIT) code CST [85] and frequency domain finite element method (FEM) code HFSS [77] are used in the numerical modeling and design. Excellent agreement between the two validates the theoretical performance of the antenna. From 172 THz to 222 THz, the designed antenna achieves return loss greater than 25 dB, as shown in Fig. 4.2. Note that the single mode bandwidth of the excited \( E_{11}^x \) waveguide mode is from 172
THz to 222 THz [106], thus directly determining the antenna operating range. The simulated
antenna end-fire gain and E-plane 0dB gain beamwidth are shown in Fig. 4.3. Good correlation

![Simulated antenna reflection coefficient with respect to the nominal port impedance.](image)

is obtained. Maximum gain is close to 13dBi near the telecom band around 195THz (1538nm)
and overall gain better than 10dBi are achieved. Gain roll-off at band edges is mainly due to the
fact that 170THz and 230THz fall out of the single-mode waveguide bandwidth. E-plane 0dBi
gain beamwidth is nearly constant around 120°, thus insuring not only wide field-of-view (as
desired for on-chip interconnect [75]), but also a very uniform and stable radiation pattern over
the whole 50THz bandwidth. The simulated radiation patterns at 175 THz, 185 THz, 205 THz,
and 220 THz are shown in Fig. 4.4. As seen, good consistency throughout the bandwidth is
observed. Note that the FIT and FEM results are very similar, and only FIT results are shown for
clarity. As seen, the co-polarized patterns in E-plane (X-Z plane) show good stability, while
those in H-plane (Y-Z plane) broaden some as frequency is increased. The cross-polarization in
Fig. 4.3 Simulated antenna end-fire gain and 0dBi E-plane beamwidth.

H-plane cannot be seen on the presented scale, while in E-plane, it increases slightly with frequency and reaches -25 dB around 220 THz.

**4.2.3 Measurement Setup**

The designed antennas are fabricated on a 200 mm diameter SOI wafer with an IMEC 193 nm deep UV lithography process [108]. A 500 nm polymer coating is deposited after the wafer is fabricated. The polymer used is a photo-definable polyimide (HD-8820) from Microsystems [110]. An SEM image of the fabricated antenna is shown in Fig. 4.5. Due to process tolerance, the structural parameters of the fabricated devices are different from the design values, both shown in Table 4.1. The fabricated length and width are 8.7% and 10.7% different than designed, respectively. The antenna tip is rounded and about 30.8% narrower than
the designed value. Note that the minimum allowed line width in the process is 120 nm, about 10% lower than our designed value. The width of the standard $220 \text{ nm} \times 450 \text{ nm}$ optical waveguide is also narrowed to 380 nm.

Fig. 4.4 Simulated (FIT) antenna radiation patterns at (a) 175 THz, (b) 185 THz, (c) 205 THz, and (d) 220 THz.
Six antenna pairs with 1 µm, 3 µm, 5 µm, 7 µm, 12 µm, and 17 µm separations are fabricated. Each pair has the antennas facing each other, so that on wafer transmission can be inferred. An SEM image of the antenna pair with 1 µm separation is shown in Fig. 4.6. All antennas are directly connected to the 220 nm × 450 nm optical waveguides, which are connected to the predefined grating couplers on the two sides of the chip [107]. The grating

<table>
<thead>
<tr>
<th>Structural Parameters</th>
<th>Fabricated Value (µm)</th>
<th>Designed Value (µm)</th>
<th>Variation percentage</th>
</tr>
</thead>
<tbody>
<tr>
<td>L3</td>
<td>2.74</td>
<td>3.00</td>
<td>8.7%</td>
</tr>
<tr>
<td>W1</td>
<td>0.38</td>
<td>0.45</td>
<td>15.6%</td>
</tr>
<tr>
<td>W2</td>
<td>0.67</td>
<td>0.75</td>
<td>10.7%</td>
</tr>
<tr>
<td>W3</td>
<td>0.09</td>
<td>0.13</td>
<td>30.8%</td>
</tr>
<tr>
<td>Antenna tip</td>
<td>Round</td>
<td>Rectangular</td>
<td>N/A</td>
</tr>
</tbody>
</table>
coupler acts as an interface between the optical waveguide and the SMF, coupling in/out light signals from the SMF. The SMF is connected to the spectrum analyzer for measurements. Note that both the spectrum analyzer and the grating coupler have narrower bandwidth than the proposed antenna, thus the measurements are limited from 190 THz to 200 THz.

The calibration is performed using a bare waveguide directly connecting two grating couplers on each side of the chip. The reference transmission level is obtained from this device. To measure the true antenna transmission, (by calibrating out the losses from the grating couplers [107] and the optical waveguides [106],) the reference transmission level is subtracted from the antenna measurements.

A super-luminescent light emitting diode (EXALOS EXS1510-2111) that has -10 dB bandwidth of 120 nm wavelength is employed as the broadband light source input. An optical spectrum analyzer, HP 71450A, is used for characterizing the transmission spectrum of these
silicon photonic devices. The sub micrometer step size and alignments of the input and output SMF are precisely controlled by two 3D micro-positioning controllers (Newport ESP301 and motors) as shown in Fig. 4.7.

![Image of measurement setup](image)

Fig. 4.7 Photograph of the measurement setup: sample is placed on the L-shape holder (black piece) at the bottom; the SMF fiber (yellow wires in the image) movements are controlled by 3D micro-positioning system; visible camera (blue cube) together with an infinity-corrected lens is used to monitor fiber movement above sample surface. (Photo courtesy of Xi.)

4.2.4 Antenna Measurement

The transmissions of the six antenna pairs without the 500nm polymer coating are measured first. Fig. 4.8 shows the obtained transmission results at 190 THz, 193 THz and 200 THz. Both the designed and fabricated antenna configurations are modeled. As seen, excellent agreement between measurements and simulations is obtained. The fabrication variations on the antenna structural parameters lower the transmissions about 3~7 dB for 7 µm and 12 µm separations, while for 1 µm separation, the transmission is better for the fabricated model. Note
that the antenna is designed to give the best performance with the 500 nm polymer coating. The measured transmissions of the polymer coated antenna pairs are shown in Fig. 4.9. The comparisons between the simulations and measurements are shown in Fig. 4.10, for 190 THz, 193 THz, and 200 THz. The actual (fabricated) antenna distance is 0.5 μm larger than designed, which lowers the antenna transmission about 1.5 dB as predicted by the corresponding simulations in Fig. 4.10. Due to the polymer coating, the light wave radiated by the antenna is confined within the polymer layer and transmitted as a two dimensional guided wave, which in turn offers better signal transmission compared to the non-coated case. As seen in Fig. 4.10,
Fig. 4.9 Measured transmissions from 190 THz to 200 THz of the six antenna pairs with different separations after polymer coating.

when the antenna separation is larger than 7 µm, the polymer coating increases the transmission between the antenna pairs by roughly 20 dB compared to the non-coated case.

4.2.5 Discussion

The transmission level comparison between antennas with and without polymer coating indicates the significance of applying an additional polymer layer in enhancing the on-chip antenna transmissions. For the dielectric rod antenna excited by the $E_{11}^x$ mode of the feeding waveguide, the E-plane is parallel to the surface of the wafer. When the wafer is coated with a polymer layer with proper thickness and refractive index, the slab $TE$ modes can be excited within the layer.

For an asymmetric slab waveguide, the cutoff frequencies of the $TE_N$ modes can be derived from [111]:
Fig. 4.10 Measured and simulated transmissions for six antenna pairs (coated) with different separations at, 190 THz (top), 193 THz (middle) and 200 THz (bottom).

\[ v_c = \tan^{-1} \left( \frac{n_3^2 - n_2^2}{n_1^2 - n_2^2} \right)^{1/2} + N\pi \]

where \( n_1, n_2, n_3 \) are the reflective indexes of the slab core, the lower substrate, and the upper cladding, respectively. \( N \) is the mode number, and \( v_c \) is the normalized cut-off frequency, which is given by:

\[ v_c = k_c a(n_1^2 - n_2^2)^{1/2} \]
where $k_c$ is the corresponding cut-off wave number and $a$ is the slab thickness. For our case, the cutoff frequencies of the $TE_0$ and $TE_1$ slab modes are 99.4 THz and 312.3 THz respectively, indicating within the antenna bandwidth region, the slab waveguide operates in a single $TE_0$ mode.

To better illustrate the power flows for the antennas with and without the polymer layer, the simulated (FIT) Poynting vectors in the antenna H-plane are shown in Fig. 4.11. With the 500 nm thick polymer coating, most of the antenna radiated power is confined and nicely guided within the polymer layer. Only a small amount of power is radiated into the BOX layer. Without the polymer layer, however, most of the radiation is directed into the BOX layer, bouncing back and forth there and eventually scattering into the silicon substrate. No power is confined on the surface of the wafer. The antenna transmission only depends on the back reflected wave from the silicon substrate in certain locations, making the transmission unstable with respect to the antenna separations, as shown in Fig. 4.8.

### 4.3 Antenna Based Signal Hub for On-chip Interconnect and Broadcast

#### 4.3.1 Configurations

To study the applicability of the proposed optical dielectric rod antenna for system level interconnections [75], optical antenna based signal hubs acting as data distribution centers are investigated. As a multi-core broadcast interconnect fabric, the hub should be highly efficient, with good output balance and channel symmetry. The antennas constituting the hub should be identical and bidirectional in that they should operate in either transmitting or receiving modes. Given these constraints, multiple antennas are specifically reshaped and arranged. Two signal hubs composed of 16 and 32 antennas are designed, fabricated and measured.
Fig. 4.11 Simulated Poynting vectors of the on-chip optical antennas with (top) and without (bottom) polymer coating.
To form the hub, the antennas are compactly arranged in a circular ring, as shown in the SEM images in Fig. 4.12. The minimum / maximum antenna distance is designed to be $g = 0.15$

Fig. 4.12 The SEM image of the fabricated signal hubs with 16 antennas (top), and 32 antennas (bottom).
µm / D = 1.422 µm and g = 0.12 µm / D = 2.8 µm for 16 and 32-channel hubs, respectively. The antenna geometry is designed to vary a little bit for each hub to give the best performance. For the 16-channel hub, W2 is shortened to be 0.45 µm, which means the antenna is directly extruded from the standard optical waveguide rather than the launcher section. The antenna length L3 is shortened to be 1.8 µm. For the 32-channel hub, W2 is also shortened to be 0.45 µm, and L3 is equal to 1.5 µm. For both hubs, the 500 nm thick polymer layer is post developed to enhance the transmissions.

### 4.3.2 Principle of Operation

The operation of the antenna based signal hubs relies on both near field coupling and far-field radiation. In that regard the transmitting antenna will couple the output signal to the nearby antennas lying within its near field region, and radiate the signal to the antennas outside this region. Take the 32-channel hub for instance. An arbitrary antenna is selected as the transmitter and labeled as Antenna 1 and the receiving antennas are labeled as 2, 3, 4, etc, accordingly, as shown in Fig. 4.12. The far-field region of the optical dielectric Antenna 1 can be estimated to be \( 2D^2 / \lambda = 3 \mu m \) away from the antenna phase center, as marked by the dashed circle. The simulations (FIT) have shown that the antenna phase center is located around the middle of the tapered section. Thus, antennas 11~23 can be considered to be within the transmitter's far-field region. The E-field intensities of the 32-channel hub obtained from the single antenna (channel) 1 excitation at 193THz are shown in Fig. 4.13. The E-field distribution clearly indicates strong coupling from Antenna 1 to Antenna 2~9 and 25~32, and a good amount of power captured by Antenna 13~21. The radiated power received by Antenna 10~12 and 22~24 are slightly weaker since these antennas are in the intermediate region. Also noticeable is the intensity of the fields
Fig. 4.13 Simulated normalized E-field intensity of the 32-channel hub (top) and isolated antenna (bottom) at 193THz (1550nm)
within the emanating spherical wavefronts in the hub are smaller than that of a stand-alone antenna, while the near-field signature within the D=2.8 µm diameter region in center is similar. This result indicates that the antennas close to the transmitter alter the radiated power distribution from the transmitter very little. That is, the scattered field from the nearby antennas has an insignificant effect on the field distribution within the hub.

### 4.3.3 Measurement

Due to the fabrication process, the dimensions of the hubs also change as shown in Table 4.2. For the 16-channel hub, the critical parameters, waveguide / antenna width W1, minimum

#### Table 4.2 Geometry Variations for 16 (top) and 32 (bottom) -Channel Signal Hubs

<table>
<thead>
<tr>
<th>Structural Parameters</th>
<th>Fabricated Value (µm)</th>
<th>Designed Value (µm)</th>
<th>Variation percentage</th>
</tr>
</thead>
<tbody>
<tr>
<td>L3</td>
<td>1.7</td>
<td>1.8</td>
<td>5.5%</td>
</tr>
<tr>
<td>W1</td>
<td>0.38</td>
<td>0.45</td>
<td>15.6%</td>
</tr>
<tr>
<td>W3</td>
<td>0.11</td>
<td>0.13</td>
<td>15.4%</td>
</tr>
<tr>
<td>Antenna tip</td>
<td>Round</td>
<td>Rectangular</td>
<td>N/A</td>
</tr>
<tr>
<td>g</td>
<td>0.15</td>
<td>0.2</td>
<td>20%</td>
</tr>
<tr>
<td>D</td>
<td>1.422</td>
<td>1.617</td>
<td>12.1%</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Structural Parameters</th>
<th>Fabricated Value (µm)</th>
<th>Designed Value (µm)</th>
<th>Variation percentage</th>
</tr>
</thead>
<tbody>
<tr>
<td>L3</td>
<td>1.425</td>
<td>1.5</td>
<td>5%</td>
</tr>
<tr>
<td>W1</td>
<td>0.44</td>
<td>0.45</td>
<td>2.2%</td>
</tr>
<tr>
<td>W3</td>
<td>0.11</td>
<td>0.13</td>
<td>15.4%</td>
</tr>
<tr>
<td>Antenna tip</td>
<td>Round</td>
<td>Rectangular</td>
<td>N/A</td>
</tr>
<tr>
<td>g</td>
<td>0.123</td>
<td>0.12</td>
<td>2.5%</td>
</tr>
<tr>
<td>D</td>
<td>2.95</td>
<td>2.8</td>
<td>5.4%</td>
</tr>
</tbody>
</table>
antenna distance $g$, and maximum antenna distance or hub radius $D$ vary more than 10%, which will inevitably impact the hub performance. For the 32-channel hub, the fabrication variations are smaller than 5.5%. It is expected that further process maturation will lead to the fabricated subsystems much closer to the designed models.

The measured and simulated (FIT) transmissions of the 16-channel hub at 190 THz, 193 THz and 200 THz are shown Fig. 4.14. As seen, the hub transmission maximum and minimum

![Graph](image.png)

Fig. 4.14 Measured and simulated (FIT) transmissions for 16-channel hub at, 190 THz (top), 193 THz (middle) and 200 THz (bottom). Results are normalized to the transmission maximum.
occur at antenna 9 and antennas 4 and 14, respectively. Antenna 9 is right in the end-fire direction of the transmitting antenna, where the maximum transmission occurs. Antennas 4 and 14 are located 68° away from the transmitting antenna end-fire direction, where the gain is < 0 dBi as predicted in the antenna study. Within this region, the antenna near field coupling starts to dominate, as the signal transmission is increasing from these two points. The hub transmission minimum, transmission imbalance (the difference between transmission maximum and minimum) and the overall transmission efficiency against frequency are shown in Fig. 4.15. In the design, the transmission minimum is maintained above -18 dB, and the imbalance is controlled to be around 5 dB. Due to the dimensional changes in the fabrication process, the measured transmission minimum is about 4 dB lower, and the imbalance is about 5 dB larger than designed. The overall efficiency is defined as the ratio of the total output power collected by all receiving antennas to the input power of the transmitting antenna. Within the measured frequency range,

![Fig. 4.15 Measured and simulated (FIT) efficiency, transmission minimum, and transmission imbalance for 16-channel hub from 190 THz to 200 THz.](image-url)
the fabricated hub has 40% to 60% overall efficiency. Compared to the designed performance, the measured efficiency is up to 20% lower (195.5 THz).

The measured and simulated (FIT) antenna transmissions of the 32-channel hub at 190 THz, 193 THz, and 200 THz are shown in Fig. 4.16. As seen, the transmission patterns from

![Graphs showing measured and simulated transmissions](image)

Fig. 4.16 Measured and simulated (FIT) transmissions for 32-channel hub at, 190 THz (top), 193 THz (middle) and 200 THz (bottom). Results are normalized to the transmission maximum.

Antenna 11 to Antenna 23 show close resemblance to the E-plane antenna radiation patterns, verifying the near-field coupling and far-field radiation mechanisms are separated around antennas 11 and 23. The transmission minimum, imbalance, and overall efficiency for the 32-
channel hub are shown in Fig. 4.17. Since the structural variations in the fabrication are better than the 16-channel hub, the transmission degradation is not as large. The transmission minimum is only about 2 dB smaller than designed. The transmission imbalance is maintained 5~6 dB throughout the measured bandwidth. However, the measured overall transmission efficiency is up to 30% smaller than designed, indicating the transmission to each receiving antenna is lower by about 2 dB. The reason is that the bare waveguide used for the calibration is from a different chip fabricated in a different IMEC run. The lithography in the two runs may have produced small geometrical parameter differences on the same device, such as the bare waveguide and the grating coupler. This in turn would cause the reference transmission level used in the calibration higher than the true level for the chip with the optical hubs, resulting in the measured transmissions about 2 dB lower than designed.

![Graph](image_url)

Fig. 4.17 Measured and simulated (FIT) efficiency, transmission minimum, and transmission imbalance for 32-channel hub from 190 THz to 200 THz.
4.3.4 Parameter Study

The overall transmission efficiency is an important measure of the broadcasting interconnection network. It is defined as the percentage of the power received by all receiving channels relative to the total available input power. Our studies have shown that the best approaches to enhance the efficiency of the wireless interconnection network are to increase field confinement and/or to decrease the distance between the transmitter and receivers.

In order to improve the field confinement, the 500nm thick polymer layer is deposited and polished over the etched SOI wafer. Shown in Fig. 4.18 are the efficiencies of the 32-channel signal hub governed by the three minimum permitted distances between antennas with and without the polymer layer. As seen, the addition of polymer almost doubled the efficiency.

![Simulated 32-channel hub transmission efficiencies for minimum antenna distances of 90nm, 120nm and 150nm, with and without polymer layer.](image)
Fig. 4.19 Simulated 32-channel hub minimum, average and maximum transmission for minimum antenna distances of 90nm, 120nm and 150nm with the top polymer layer (top) and without the top polymer layer (bottom).
over the entire bandwidth. More importantly, the polymer dramatically increased the minimum hub transmission (see Fig. 4.19). Throughout the signal hub bandwidth, the polymer enhanced the minimum transmission for 10 to 20 dB when compared to the case without the polymer.

The minimum distance between the antennas is limited by the fabrication rules. For example, IMEC 193nm deep UV lithography specifies the minimal distance of 100nm. The baseline signal hub is built with a 120nm minimum distance thus easily satisfying this criterion. To assess the role of the fabrication rules, 90nm, 120nm and 150nm minimal distances between antennas are used. Fig. 4.18 and 4.19 depict the effects over the whole bandwidth. As seen, as the antenna distance decreases, the transmission efficiency increase correspondingly. However, the positive effects on the transmission minima are not noticeable. The decrease in distance actually shifts the hub response to higher frequencies.

The number of channels is another important parameter for the optical signal hub. Fig. 4.20 shows the simulated overall efficiencies and transmission minima for 32, 16 and 8 channel hubs. Note that due to the computer memory limitations, the hub configurations with more than 32 channels have not been studied. For each case, the antenna geometrical parameters are tuned to give the best performance in terms of efficiency and transmission. The minimal antenna distance is kept above 120nm for fabrication consideration. As seen, the overall transmission efficiency increases as the channel number increases indicating, as expected, that with a larger antenna count, more of the incident power will be captured. The hub bandwidth on the other hand is virtually unaffected by the antenna number thus favoring future increases in channel count. For 16 channel hub, simulated results show the transmission minima has around 2.5dB improvement compared to that of the 32 channel hub. However, for 8-channel hub, the transmission minima is close to that of the 16 channels case. This trend is due to the limited
transmission via near field coupling to about 17dB if the antennas are sparsely distributed, especially for the first two receiving antennas in immediate neighborhood of the transmitter.

The thickness and permittivity of the coating polymer layer may also vary. Three different layer thicknesses and permittivities are chosen to evaluate the corresponding effects on the 32-channel hub performance. Obtained results are shown in Fig. 4.21. As with the IMEC process variations, the polymer has little effect on the overall performance of the signal hub.

4.3.5 Microwave Counterpart

To further validate the hub operating theory, an 8-channel microwave wireless interconnection network is designed, fabricated and measured. Microstrip fed Vivaldi antennas [112] are used as transmitters and receivers. Note that the actual scaling of the dielectric rods and
SOI wafer is not practical, however, the basic physics of the optical signal hub is properly captured by the chosen antennas and fabricated layout. Antennas and feeders are etched on a 60mils thick FR4. Field confinement is emulated by addition of an FR4 disk on the top-side of the antenna layer.

Fabricated microwave signal hubs are shown in Fig. 4.22. As seen, 2.5cm long Vivaldi antennas have 1.2cm wide flare opening. They are fed by the 50Ω microstrip lines with wideband microstrip to slot line transitions. Antennas are compactly arranged in a d2=3.4cm circular ring thus achieving good coupling. For measurements, the unused SMA ports were terminated with 50Ω loads.

Measured and simulated efficiencies are shown in Fig. 4.23, while the spread of transmissions is shown in Fig. 4.24. As seen, excellent agreement between the theory and
Fig. 4.22 Fabricated microwave signal hub. (left) layout, (right) photograph.

Fig. 4.23 Measured and simulated (FIT) microwave interconnect network efficiency. FR4 has permittivity and loss tangent of 4 and 0.015, respectively.
measurements is obtained. Interconnect has 40~50% efficiency within the 3GHz to 4GHz bandwidth, which is about 5% smaller than its optical counterpart. The transmission minima is about -15dB, slightly better than its optical counterpart. Even though the microwave hub is not a true scaled version of the optical hub, the obtained results clearly show that the theory and design at optical frequencies are valid.

4.4 Conclusion

Nanoscale wideband on-chip optical dielectric rod antennas and signal hubs are demonstrated in this chapter. The designed optical antennas have return loss above 25 dB, gain greater than 9 dBi, and consistent radiation patterns from 172 THz to 222 THz. This demonstrated 50THz bandwidth is to the best of our knowledge the widest instantaneous bandwidth to date. Six optical antenna pairs are fabricated and on-chip antenna transmissions are
measured from 190 THz to 200 THz. The obtained results show good agreement with numerical modeling. Significantly improved transmissions are observed when a 500 nm thick polymer layer is deposited on the wafer to confine the antenna radiation. Two on-chip optical hubs consist of 16 and 32 antennas are also designed, fabricated and measured. -22 dB transmission minimum, 5 dB transmission imbalance, and over 40% overall efficiency are measured for the 32-channel hub. Measured results also indicate that the proposed optical hub fully satisfy the WDM requirement in the core-to-core interconnections for the future multi-core microprocessors.

Compared to the wideband antennas discussed in Chapter 2 and 3, the optical dielectric rod antenna only achieves 25.4% fractional bandwidth. However, its absolute bandwidth is 50THz, 4 and 3 orders higher than the microwave patch and the MMW LPDA antennas, respectively. For on-chip WDM signal transmission, this bandwidth provides significantly higher data throughput compared to the traditional electrical interconnects. Compared with the optical plasmonic antennas, which use nano-scale metallic structures as radiators, the antenna proposed in this chapter employs dielectric material as its feed and radiator. Much higher radiation efficiency and bandwidth are achieved with characterizable transmission performance between antennas.
CHAPTER V

SUMMARY, CONTRIBUTIONS AND FUTURE WORK

The research presented in this thesis has demonstrated several new antenna designs and concepts that can be used in emerging commercial and military applications requiring ever increasing instantaneous system bandwidth. Research covers L/S band, Ku through W-band, and infrared spectrum. A traditional narrowband antenna bandwidth enhancement, extension of performance limits, and wireless transmission beyond traditional bands are demonstrated.

5.1 Microwave Wideband Patch Antennas

Chapter 2 proposes a novel feeding method to increase the bandwidth of microwave patch antennas. Two cylindrical differentially-fed probes are used to capacitively excite the patch antenna. Over 60% impedance and far-field bandwidths are obtained, way higher than the previously demonstrated patches.

The basic idea originated from the L- and T-probe fed patches, where a second resonance is used to increase their bandwidths. However, these approaches achieve only up to 40% impedance bandwidth and suffer from weak mechanical robustness and radiation pattern distortion caused by the higher order mode radiation at the upper portion of the impedance bandwidth. The thesis proposes cylindrical probe feeding to reduce the inductance from the feeding probe and lower the Q-factor of the patch, which in turn increases the impedance
bandwidth. However, if a single probe feed is used, the antenna still operates under higher order mode which renders pattern distortion at high frequencies. Differential-fed approach developed in this thesis enforces the radiations from the current flowing toward the non-radiating edges to cancel out in the far-field, thus reducing the high cross-polarization in H-plane. An antenna prototype designed based on the differential cylindrical probe feed approach, referred as Antenna 1, is fabricated and tested. Excellent VSWR and gain bandwidths are obtained (103% and 68%, respectively). Nonetheless, when two probes are used, the mutual coupling between the two inputs is significant and Antenna 1 suffers from excessive loss at lower frequencies. To address this issue, a shorting wall is introduced between the two probes connecting the patch to the ground plane redirecting the current flowing. The modified patch configuration referred as Antenna 2 has mutual coupling < -11dB across the entire bandwidth. About 60% impedance and far-field bandwidths are achieved respectively with 56%/48% overall bandwidth for VSWR<2 and gain>5dBi/8dBi.

5.2 Millimeter-Wave Log-Periodic Antennas

In millimeter-wave frequencies, wideband antenna development is mainly constrained by limited fabrication capabilities. For instance, LPDA antenna requires an integrated transmission line for proper feeding. Traditional antenna fabrication methods, such as PCB, have various issues when millimeter-wave LPDA antenna realization is desired. The rapid advance in photolithograph technologies has enabled development of new miniature millimeter-wave devices. Chapter 3 demonstrates that millimeter-wave LPDA antennas with monolithically integrated passive components are now readily achievable using the newly developed fabrication technologies.
The PolyStrata process used in the antenna fabrication in this thesis can easily combine more than 10 strata layers with auxiliary dielectric straps supporting suspended metallic structures, such as the center conductor of a coaxial line. However, the layer thicknesses in different runs are not consistent and vary between 20µm and 125µm. To provide a general design guideline not only limited to PolyStrata but for other current or future thick photo-lithography micromachining based millimeter-wave LPDA antennas, constant 100µm thick layers are used in the design process. Up to 110GHz operation can be easily achieved with properly designed antenna boom and dipole configurations. Above 110GHz, the antennas suffer from the increased cross-polarized radiations from the electrically large boom and the cylindrical post connecting the coaxial center conductor and the upperboom. To reject the 60GHz unlicensed band, the straight 50Ω recta-coaxial line feed in the antenna's low boom is redesigned based on a band rejection filter topology.

By implementing the specific strata configurations of two different runs in the generalized LPDA model, two LPDA antennas are fabricated and tested. VSWR<2.5 is obtained for both antennas within the tested bandwidth. Far-field measurements performed on fabricated antenna show excellent correlation with simulations. Consistent patterns, E and H plane 3dB beamwidth, and 10dBi directivity are measured throughout the bandwidth. For dual-linear and dual-circular polarizations not easily achievable with LPDA antennas, millimeter-wave planar log-periodic antennas are designed and fabricated. A PCB process, which allows 3mils minimal trace width and lower costs is used. The antennas are composed of two boards which are carefully aligned and attached with dowel pins and nylon screws. VSWR<3 and good far-field radiation consistency are measured within the 18GHz to 40GHz bandwidth.
5.3 Optical Antenna for On-chip Communication

Chapter 4 discusses the development of on-chip optical antennas for the next generation microprocessor core-to-core interconnect. Optical dielectric antennas and signal hubs are designed on SOI wafer for signal exchanging and broadcasting via on-chip wireless links formed over a 500nm thick polymer layer. 50THz bandwidth, the widest known to date, is achieved for the designed devices. Transmission measurements performed from 190THz to 200THz validate the predictions made by computational methods.

The dielectric optical antennas are designed to overcome loss and fabrication challenges when metallic traces are used at these frequencies. The antennas are built on the top silicon layer of the 200mm SOI wafer. For desired radiation, an exponentially tapered antenna shape is used, which is fed from the standard single mode optical waveguide via a slightly wider beam. To prevent the antenna radiated power scattering into the substrates underneath, a 500nm thick polymer layer is deposited on top of the wafer to conform the radiated power within. Antenna transmission measurements performed on 6 antenna pairs with different tip-to-tip separations show excellent agreement with the theoretical results. For on-chip signal broadcast, optical signal hubs are proposed, which consist of carefully placed antennas in a circular ring. 16- and 32-channel hubs are fabricated and measured. -22.4dB transmission minimum, 5dB output imbalance and 45% overall efficiency are measured for the 32-channel signal hub. With a 50THz bandwidth, this hub would offer significantly higher data throughput with much lesser signal latency and power consumptions compared to the traditional electrical interconnect solutions.

5.4 Original Contributions

The original contributions of this thesis are as follows:
• Developed a cylindrical probe feeding method to increase the bandwidth of the microwave patch antennas and maintain their mechanical robustness.

• Implemented differential feeding to alter the surface current flow toward the none radiation edges of the patch antenna to reduce the radiation pattern distortion at the higher portion of the impedance bandwidth.

• Demonstrated that a metallic wall inserted between the differential feed probes and shorting the patch to the ground plane reduces the losses produced by the mutual coupling.

• Demonstrated that the higher sidelobes in E-plane of the patch antenna at high in band frequencies are caused by the increased electrical distance between the two radiating edges.

• Demonstrated that the design and fabrication of high performance millimeter-wave LPDA antennas are now feasible using the newly developed thick photo-lithography technologies.

• Developed a general design guideline for current and future thick photo-lithography micromachining based millimeter-wave LPDA antennas.

• Developed a band rejection feed that can be monolithically integrated with the LPDA antenna for signal rejection at specific frequencies.

• Developed a low-cost dual-polarized millimeter-wave planar log-periodic antenna compatible with PCB manufacturing process.

• Developed on-chip optical antennas for next generation microprocessor core-to-core interconnection.
• Demonstrated the importance of the polymer coating in increasing the transmission between on-chip optical antennas.

• Developed on-chip optical signal hubs based on optical antennas for data broadcasting. Established a novel operating principle to increase the hub efficiency while reduce the output power imbalance.

5.5 Future work

The research described in this thesis can be expanded in many different ways. The following are only a few possible directions.

5.5.1 Microwave Wideband Patch Antennas

The proposed patch antennas in Chapter 2 exhibit large sidelobes at high in band frequencies. One method to reduce the sidelobes is to use curved or zigzagged patch configurations, which can effectively shorten the electrical distance between the two radiating edges. However, the impact from the exotic patch shaping on the other antenna performance parameters needs to be studied and minimized.

The proposed patch antennas are linearly polarized. For many applications, it is desired to design circularly polarized low-cost wideband patch antennas. Patch shaping, impact of coupling and methods for mitigation thereof as well as feeding approach need to be researched to establish the simple or dual wideband circularly polarized operation.

The scalability of the proposed patch antenna to both lower and higher regions is also worth further investigation. By using other fabrication techniques, such as the one used in the LPDA antenna fabrication, millimeter-wave wideband patch is possibly achievable. The
theoretical studies shown in the Appendix A have been conducted, however, the experimental validation still needs to be performed.

5.5.2 Millimeter-Wave Log-Periodic Antennas

A possible further research in millimeter-wave LPDA antennas is in their arrays. By deploying multiple LPDA antennas in an array configuration, beam control can be achieved by using beamforming networks. The study on a W-band two-element LPDA antenna array is shown in Appendix B, however, experimental investigation is needed.

A millimeter-wave planar log-periodic antenna is designed and fabricated using PCB process in Chapter 3. However, without a wideband quadrature hybrid, the antenna only operates under dual-linear polarization mode. A wideband millimeter-wave quadrature hybrid should be designed compatible with the PCB process for monolithic integration with the antenna. Due to the thick prepreg used in the standard multilayer PCB bonding, the antenna electrical performance will change when the two antenna boards are joined together. The screw bonding method used in the thesis also alters the antenna far-field performance. An improved bonding method should be investigated to maintain the antenna performance and increase the antenna robustness for potential conformal mounting on curved surfaces.

5.5.3 Optical Antenna for On-chip Communication

The optical dielectric antennas developed in Chapter 4 are highly directional. Even though direct transmission between two oppositely placed antennas is measured, a full 360° transmission measurement around a single antenna is still preferred in order to characterize the antenna far-field performance.
The signal hub is designed with multiple antennas compactly arranged in a circular ring using both near field coupling and far-field radiation for signal broadcast. Although the demonstrated 32-channel hub shows excellent performance, when the number of channels is increased above 64, due to the increased ring diameter, the overall hub efficiency drops and the output imbalance degrades. This issue needs to be addressed by developing robust methods that will enable additional channels into the hub. For example, by placing some auxiliary scatterers in the middle of the ring, an increase in the near field coupling can be achieved.

An omnidirectional on-chip optical antenna can provide a different concept to the hub. A theoretical antenna design is proposed in Appendix D, which achieves 240º 3dB beamwidth at 205THz. However, the 3dB beamwidth gets lower at lower frequencies. At 185THz, the 3dB beamwidth is decreased to 180º. To get stable omnidirectional radiation characteristic over a wide frequency range, the antenna design and radiation direction control still require further investigation.
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APPENDIX A: WIDEBAND W-BAND PATCH ANTENNA

A.1 Introduction

Rapid progress in micro-fabrication technologies has initiated intensive study and development of various miniature millimeter-wave (mmW) devices. For example, in [113], a miniature 310µm tall inhomogeneous copper recta-coax lines and related components are fabricated using PolyStrata, developed by the researchers from Nuvotronics, LLC, BAE Systems and University of Colorado. In [114], Ka-band quasi-planar high-Q resonators are developed using the same process. Micro-fabricated patch antennas and arrays are also demonstrated at Ka-band [115-117], V-band [118] and W-band [119]. It is seen that patch antennas scaling to mmWs worsens their inherently narrow-band nature, primarily due to the difficulties in manufacturing and use of traditional bandwidth extension techniques. The antennas listed above have bandwidths below 10%. In spite of bandwidth enhancement to about 13.3% [120], the narrow bandwidth still tempers the use of patch antennas for emerging wideband mmW applications.

A PolyStrata based wideband W-band patch antenna is presented in this appendix. Following the bandwidth extension approach from Chapter 2, the antenna is designed to feature dual circular, displacement currents coupled differential feed. A vertical metallic wall is added to enhance the isolation between the probes and to improve the mechanical rigidity of the patch. The developed antenna is cavity backed, and monolithically integrated with a wideband micro-coaxial 180° hybrid (designed as combination of a 90° coupled line hybrid and 90° phase shifter),
which provides the necessary differential excitation. Finite integration technique (FIT) code CST-MWS [85], and finite element method (FEM) code Ansys-HFSS [77] are used in the design. The developed antenna has operating bandwidth from 76.5GHz to 99.5GHz, and from 75GHz to 120GHz for VSWR<2 and 2.5, respectively. The bandwidth for broadside realized gain above 7dBi is from 75GHz to 115GHz.

A.2 Antenna Geometry

The antenna is designed and fabricated using the same photo-lithography technique as in Chapter 3, PolyStrata. The configuration of the developed antenna and the photograph of the fabricated device are shown in Fig. A.1 (a) and (b), respectively. As seen in Fig. A.1 (a), the antenna consists of two parts; part A, the cavity-backed patch, and part B, the 180° hybrid, which is designed as a cascade of a 90° hybrid and two 45° Schiffman phase shifters. For part A, the cavity is 3.5mm × 3.5mm, while the patch is 1.25mm × 0.9mm. A1 and A2 are the differential ports connected to the two ports of the 180° hybrid. For part B, the hybrid is 9mm long and 6.7mm wide. B1 is the differential input port, while B2 is the sum input port, which is the isolated port for this configuration. The overall height of the designed antenna is 0.825mm, and is driven by the given strata configuration. Note that the further strata optimization and selection can yield even better performance of the wideband patch antennas.

The two antenna feeding probes are directly connected to the inner conductors of the two output ports of the 180° hybrid at the back side of the antenna, as seen from Fig. 2 (a). Two metallic posts are fabricated at both side of each probe acting as mechanical holders, thus allowing the landing of the supporting dielectric strap. The two circular probes have 0.45mm
Fig. A.1 The 3-D model of the antenna configuration (a) and the photograph of the fabricated device (b).
diameter and are 1.3mm separated from each other. The air gap between the probe and the patch is 0.1mm, providing the necessary capacitance for the probe resonance. The supporting wall and the wall of the cavity are 0.2mm thick. The antenna cavity is designed mainly for facilitating the flowing of the photo resist, with minimal electrical contributions.

A.3 Antenna Performance

The antenna analysis and design are conducted using the time domain finite integration technique (FIT) code CST-MWS and frequency domain finite element method (FEM) code Ansys-HFSS. The cavity backed patch antenna, part A and the 180° hybrid, part B are designed separately and integrated upon completion of their individual designs. The FIT computed impedance performance of the antenna part A is shown in Fig. A.2. As seen, the reflection

![Simulated (FIT) S-parameter of the antenna part A.](image)

Fig. A.2 Simulated (FIT) S-parameter of the antenna part A.
coefficient of the two antenna input ports A1 and A2 are below -10dB from 70.5GHz to 120GHz. The port coupling is below -12dB throughout this bandwidth. The antenna is designed to be symmetric.

The FEM performance of the hybrid is shown in Fig. A.3 and Fig. A.4. For these two figures ports 1, 2, 3, and 4 are assigned to B2, A1, A2, and B1 of Fig. A.1 (a), respectively. From Fig. A.3 the reflection coefficient is below -10dB for all frequencies above 68GHz. The quadrature hybrid itself is symmetric; however, the phase shifters are not resulting in different reflection coefficients for all ports. Magnitude and phase imbalances are shown in Fig. A.4 for both the sum and difference input ports. Ideally the output ports (2 and 3) should have equal s-parameter magnitude using either input port. It can be seen that for this component the output magnitudes are within 0.8dB over the entire frequency range. Almost all of this imbalance is from the coupled line 90° hybrid and can be reduced by integrating additional quarter-wave
sections at the expense of increased loss. The phase shifters contribute minimal magnitude imbalance due to different line loss between the coupled line and delay line paths. The phase imbalance of this component is $0^\circ \pm 6^\circ$ and $180^\circ \pm 5^\circ$ for the sum and difference inputs, respectively, across the entire band. The quadrature hybrid has inherently low phase misbalance and this parameter can be controlled through the phase shifter design. For this design two $45^\circ$ phase shifters are cascaded to provide less phase imbalance than a $90^\circ$ phase shifter with the same number of coupled line sections. The phase shifter phase imbalance can be reduced by increasing the number of coupled line sections or by reducing the stage phase shift. Much like the hybrid reducing the imbalance of the phase shifter comes at the expense of more loss.

The computed S-parameters of the overall developed antenna are shown in Fig. A.5. Ports B1 and B2 are denoted as 1 and 2 for the S matrix. As seen, the designed antenna operates from 76.5GHz to 99.5GHz with $S11<-10\text{dB}$ (VSWR<2), which is a 26% impedance bandwidth.
Fig. A.5 Computed S-parameters of the W-band patch.

Fig. A.6 Computed realized broadside gain of the W-band patch.
For $S_{11}<-7.3\text{dB}$ (VSWR$<2.5$), the impedance bandwidth is from 75GHz to 120GHz. The port isolation is better than 10dB from 62GHz to 119GHz. Note that the ripple shown on $S_{11}$ from 99.5GHz to 102.5GHz is caused by the integration of the patch and the hybrid, where the port couplings of the two devices stack up. The antenna realized gain at broadside is above 7dBi from 75GHz to 115GHz as shown in Fig. A.6. The far-field radiation patterns at 80GHz and 95GHz are shown in Fig. A.7. Some squinting in the E-plane is the consequence of the necessity for the

![Fig. A.7 Radiation patterns at 80GHz: (a) E-plane, (b) H-plane, and 95GHz: (c) E-plane, (d) H-plane](image-url)
design to follow the strict fabrication requirements in terms of strata configuration and aspect ratios. Thus, these can be improved in the future with a process tailored to these devices. Note that the FEM and FIT give excellent agreements.

A.4 Conclusion

Appendix A demonstrated a wideband W-band patch antenna with monolithically integrated 180° hybrid. The design follows the fabrication requirement of a rapidly maturing sequential micromachining process PolyStrata. Capacitive differential probe feeding is used to increase the antenna bandwidth. Copper wall is used between the two probes to reduce the port coupling and support the rectangular patch while the cavity backing is added to easy some of the fabrication requirements. Impedance bandwidth is from 76.5GHz to 99.5GHz and from 75GHz to 120GHz for S11<- 10dB (VSWR<2) and S11<-7.3dB (VSWR<2.5) respectively. The port isolation is better than 10dB from 62GHz to 119GHz. Broadside realized gain is above 7dBi from 75GHz to 115GHz. Predications are verified using time domain finite integration technique (FIT) code CST-MWS and frequency domain finite element method (FEM) code Ansys-HFSS.
APPENDIX B: W-BAND END-FIRE LOG PERIODIC DIPOLE ARRAY

B.1 Introduction

Appendix B demonstrates a W-band LPDA antenna and a two element V through W-band LPDA antenna array. They are also designed for and fabricated with PolyStrata. Total of 13 dipoles are crisscrossed in two separated copper layers occupying overall volume of 7mm × 2.6mm × 0.915mm. A monolithically integrated μ-coaxial line in the lower boom is used as the antenna feed. Two quarter wavelength shorting stubs are connected to the feeding line at the low frequency end of the antenna to mechanically support the inner connector of the μ-coaxial line. Antenna achieves return loss above 10dB and end-fire gain greater than 7dBi from 70GHz to 115GHz. Consistent radiation patterns throughout the bandwidth are also observed. A two element LPDA based array monolithically integrated with a wideband 180° hybrid is also designed and fabricated. Previously designed LPDA antenna is slightly modified to operate from 50GHz to 110GHz, thus spanning two traditional rectangular waveguide bands within a single non-dispersive and low-loss transmission line technology. Hybrid is designed as a combination of a wideband 90° coupled line hybrid and two 45° Schiffman phase shifters. Good end-fire gain for the sum mode and deep null for the difference mode are achieved. The performance of the developed antenna and array is validated using full wave analysis based on the time domain finite integration technique code CST-MWS [85] and frequency domain finite element method code Ansys HFSS [77].
B.2 Antenna Geometry

The LPDA antenna is fabricated using the sequential surface micromachining process, PolyStrata, developed by the BAE Systems, Nuvotronics and University of Colorado, under the DARPA-MTO funded program 3-D MERFS [121]. A scanning electron microscope (SEM) image of the fabricated antenna is shown in Fig. B.1. The antenna is built from 11 copper layers sequentially deposited on a silicon wafer (used as a carrier substrate that the antenna will be detached from once the fabrication is completed). The preset layer thicknesses vary from 20\(\mu\)m to 125\(\mu\)m. Note that the layers (strata) composition was not optimized for this design, but it was chosen to accommodate the fabrication of many different components on the same fabrication run. The overall volume occupied by the antenna is 7mm \(\times\) 2.6mm \(\times\) 0.915mm. Detailed
drawings showing the antenna structure, location of the feeding line, applied method to connect the feeder to the antenna and to hold the inner conductor in place, are depicted in Fig. B.2. As seen, the feeding 50\,\Omega \, \mu\text{-coaxial line is integrated in a 0.54mm thick lower boom. The boom thickness is determined by the preset first five layer heights and its width is linearly tapered from 0.59\,mm to 0.5\,mm. Note that a sequential build of a recta-coax line requires a minimum of 5 strata. The outer and inner conductors of the feeding \( \mu\text{-coaxial line have fixed heights of 0.2\,mm and 0.05\,mm, respectively. The widths are changing from 0.37\,mm to 0.3\,mm and from 0.14\,mm to 0.1\,mm, respectively, to ensure 50\,\Omega \) line impedance. The openings in the side walls of the recta-coax feeder line, herein referred to as release holes, are used for the removal of the
sacrificial photoresist. Although they are placed periodically, any distribution is fine as long as the sufficient number of the holes is present so that the appropriate flow of the chemical etchants is permitted. The full-wave simulations have shown that their effect is minimal. The antenna upper boom is solid with the same width as the lower boom, with a thickness fixed at 0.25mm as determined by the corresponding layers. The separation between the lower and upper booms is 0.125mm. A total of 13 dipoles are used to form the log periodic array. The longest dipole is 2.6mm × 0.32mm, while the shortest one is 0.77mm × 0.1mm (boom width included). Note that the antenna upper dipole arms are thicker (100µm) than the lower ones (50µm), due to the fixed/preset layer thicknesses in the fabrication. The growth rate τ of the log periodic dipoles exponentially decays from 0.91 to 0.89 from the longest to the smallest dipole to mitigate the effects of the increased reflection coefficient due to the thick boom. The spacing factor σ is constant at 0.2. The dipole arm width and length ratio is chosen to be 1:8 to achieve the best electrical performance while maintaining good mechanical rigidity. The upper and lower booms of the antenna are physically connected at 0.5mm away from the longest dipole to enable necessary mechanical support. A fat cylinder at the end of the feeder line is used to connect the upper boom to the inner conductor of the lower boom, as shown in Fig. B.2. The inner conductor of the µ-coaxial line in the lower boom is mechanically supported by the two quarter wave length shorting stubs (1mm × 0.05mm). The two shorting stubs work as a bandpass filter with passband (S11<-10dB) from 70GHz to 120GHz, so that the impedance performance of the antenna is least affected.

B.3 Antenna Performance
The antenna analysis and design are conducted using the time domain finite integration technique (FIT) code CST-MWS [77] and frequency domain finite element method (FEM) code Ansys-HFSS [85]. The analysis included the determination of near- and far-field effects of growth rate, teeth lengths and widths, mechanical support, quarter-wave stubs, cylinder transition, etc. for the preset strata heights. The antenna reflection coefficient and endfire realized gain are shown in Fig. B.3. As seen, the antenna achieves return loss above 10dB and realized gain greater than 7dBi from 70GHz to 115GHz. Good correlation between the two different numerical methods is also observed, thus fully verifying its theoretical performance. Note that the antenna also has front to back ratio greater than 15dB.

The antenna radiation patterns at 80GHz and 100GHz are shown in Fig. B.4 and Fig. B.5, respectively. Throughout the design process, the consistency of the radiation patterns was one of the main criteria for determining and accepting its far-field performance. The 3dB beamwidth in both planes, beam squint, gain, front-to-back ratio, and cross-polarizations were the electrical
measures used to determine the pattern consistency. As seen from Fig. B.4 and Fig. B.5, the E-plane co-polarization beam maximum is 10° off the end-fire direction. The full-wave simulations have shown this is due to the fabrication (preset strata) driven asymmetries in dipole
arrangements, the necessity to integrate the boom with the feeder coax, and the overall height to wavelength ratio between the feeder and the operating wavelength. Note that in the previous LPDAs, the operating wavelength was much larger than any vertical dimension of the antenna so that the boom/feeder effects are negligible. Also seen in the figures is that the antenna cross polarization is 10dB below the maximum gain. This level of cross-polarization is larger than with typical low-frequency LPDAs and it is caused by the large boom thickness and the length of the cylinder feed attachment (denoted in Fig. B.2 as Cylinder Balun). This feed increases the cross polarization due to its dipole like radiation which in E-plane is orthogonal to that of the LPDA. Note that better performance would be expected for the process tailored specifically and exclusively for the LPDAs. Finally, we note that the location of the shielding of the quarter wavelengths stubs is chosen so that the effect on the antenna near-and far-field is kept at minimum.

**B.4 LPDA-Based Two-Element Array**

Two LPDA antennas are modified first and then monolithically integrated with a 180° hybrid to form a two element array with sum and differential operating modes. A 3-D model of the designed array is shown in Fig. B.6. As seen, to reduce the antenna dimension, the two shorting stubs used to support the inner conductor of the μ-coaxial line are removed. 13 dielectric straps are instead used as a mechanical support. The dipole length at the low frequency end is increased to 3.4mm and the bandwidth is from 50GHz to 110GHz with S11<-10dB and gain>7dBi. Antennas are tilted 18° inward to enhance common mode gain and the depth of the differential mode null. Distance between the dipoles of the low frequency end is 1.3mm. 180° hybrid is designed as a cascade of a 90° hybrid and two 45° Schiffman phase shifters. The
Fig. B.6 Drawing of the LPDA-based array with integrated 180° hybrid. (backside view with bottom layer removed)

overall dimension of the array embodiment is approximately 16mm by 6mm. Note that the CPW launches used for the measurement are also included.

Array radiation patterns for sum and differential inputs at 50GHz are shown in Fig. B.7. As seen, the sum mode antenna is 10dBi and cross polarization level is <-20dB. For differential input, the antenna achieves -15dBi null at end-fire direction. At higher frequencies, this number is decreased slightly due to the hybrid imbalance and increased cross polarization.
Fig. B.7 Simulated (FIT) LPDA antenna array radiation patterns at 50GHz. (a) common mode, (b) differential mode.

B.5 Conclusion

Appendix B demonstrated a W-band LPDA antenna and a two element V+W band LPDA-based array. Antennas are designed for and fabricated with 11 strata PolyStrata process. A μ-coaxial line is monolithically integrated in the antenna lower boom as the feeding line. Return loss above 10dB, end-fire realized gain greater than 7dBi, and consistent patterns are achieved from 70GHz to 115GHz. An LPDA-based two-element array operating from 50GHz to 110GHz is also designed and fabricated. Good end-fire direction gain and null are achieved for corresponding operating modes.
APPENDIX C: ON-CHIP OPTICAL SIGNAL HUB INTEGRATION ON A MULTI-CORE CHIP (COURTESY OF ZHENG LI)

Appendix C discusses the conceptual optical signal hub on-chip integration for emerging many-core ICs. As shown in Fig. C.1, three-dimensional (3D) heterogeneous IC integration technology is considered. The nanophotonic network is in the top layer which is composed of the optical signal hub and other nanophotonic components including transmitters, receivers, channel waveguides, WDM filters and etc. The signal hub is located in the center, broadcasting and receiving optical signals. Leveraging WDM, multiple signal streams can be broadcast concurrently through different wavelength channels. The antenna feeds, i.e. photonic waveguides, extend to distributed on-chip photonic transceivers that relay messages among on-chip processor cores. Photonic link transceiver designs using micro-ring resonator-enhanced silicon modulators
and waveguide-integrated Germanium-doped silicon photodetectors have been demonstrated to be compact, low power, and able to deliver over gigahertz speed [122]. The bottom layer shown in Fig. C.1 hosts CMOS amplifiers and other electrical components. Next, the power and performance benefits of optical signal hub based network design are estimated. Detailed analysis and experimental studies from many-core system perspective are presented in [123].

Performance wise, the signal hub provides light-speed broadband transmission for multi-core systems. The overall communication bandwidth equals the number of WDM channels times the bit rate per channel. The number of WDM channels is in turn determined by the bandwidth of the optical signal hub. Assuming the 100GHz ITU grid in C+L band where silicon photonic devices are usually operating, the 50THz (400nm) bandwidth working range of the hub could in principle offer hundreds of WDM channels. If we further assume that each on-chip processor core issues one broadcast message every 10 clock cycles, these WDM channels can support over 100 processor cores broadcasting simultaneously, thereby providing sufficient communication bandwidth for emerging multi-core ICs. Note that the proposed optical signal hub can maintain the desired communication bandwidth, which is nearly independent of the on-chip processor core counts. A 32-channel hub described in the Chapter 3 easily achieves the targeted 50THz bandwidth.

The power efficiency of the designed optical signal hub can be estimated by computing the loss budget [124], which demonstrates that the hub can operate reliably in the mW power range, thus significantly outperforming alternative electrical or other photonic interconnects. Using the loss budget analysis method, the communication system components are quantified using a series of transmission coefficients. The attenuated power arrived at the receiver end
should provide enough signal-to-noise ratio so that the targeted bit-error-rate (BER) at multi-gigabit per second transmission could be reached.

More specifically, a BER of $10^{-12}$ is typically required for multi-gigabit transmission. Following the derivation from [125], the BER can be determined using Gaussian noise model as follows:

$$BER = \frac{1}{2} \text{erfc}(Q/\sqrt{2}), \quad Q = \frac{I_1 - I_0}{\sigma_1 + \sigma_0}$$

where \(\text{erfc}\) is the complementary error function, \(I_1\) and \(I_0\) are the signal “1” and “0”s corresponding current, and \(\sigma_1, \sigma_0\) are the corresponding noise standard deviations.

Optical power is converted to electrical signal power in the photodetector with responsivity of \(R\). Responsivity is the ratio of the generated photocurrent to the incident optical power, which is typically 1A/W. Noise power is dominated by thermal noise in multi-gigabit transmission. Thus, the minimal required optical power at the receiver’s end, i.e. receiver sensitivity, can be determined as follows:

$$P_{\text{optical}} = \frac{2Q \sqrt{P_{\text{NA}}}}{R \sqrt{Z_L}}$$

where \(P_{\text{NA}}\) is the thermal noise power, which is equal to 0.2nW at room temperature. \(Z_L\) is the load resistance. Appropriate circuit design can be leveraged to optimize receiver sensitivity by setting appropriate bandwidth and load resistance. Theoretically, for multi-gigabit transmission \(P_{\text{optical}}\) could reach -44dBm. Recent research has demonstrated that this limit is possible to attain. For example, Beaisoleil et al. predicted sensitivity below 130nW [74], while a photodetector with 200nW measured sensitivity is demonstrated in [126].

The fundamental parameter describing attenuation in the proposed broadcast system is the power transmission factor \(T\), with the worst case of -22.4dB. Considering the power losses of
the other communication components, optical couplers used to couple off-chip fiber to on-chip waveguide have ~3dB loss [127]. Modulators and filters in transmitter end have ~2dB loss [128]. Waveguides have 1.34dB/cm loss [129]. De-multiplexers in the receiver end have 1.8dB loss [130]. The analysis above shows that, considering the power losses of all the communication components, to achieve the required BER, a power budget of 920 µW per channel is sufficient, assuming a conservative -30dBm receiver sensitivity. Supporting the 32 WDM channels thus requires approximately 37mW, which is orders of magnitude more efficient than electrical interconnect alternatives.
The optical antennas introduced in Chapter 4 are highly directional and are good for signal transmission and reception in a specific direction. However, for signal broadcast, omnidirectional antennas, such as dipoles or monopoles, are preferred. As discussed in Chapter 4, metallic antennas are not suitable for optical signal radiation due to various reasons. To design an optical dielectric antenna with omnidirectional radiation patterns becomes a new challenge. In on-wafer environment, the design of nanoscale dielectric antenna is high limited by manufacturing capabilities. A finite thickness 2D structure can be fabricated with restricted minimal trace and gap widths. A true nanoscale 3D structure, however, is a big challenge to the current fabrication technologies. To achieve an omnidirectional radiation using 2D patterns filled with dielectric materials requires unconventional designs and novel antenna geometries.

Since dielectric antennas need to be fed from a waveguide in the same silicon layer, a pure omnidirectional pattern is not likely achievable due to the wave scattering from the feeding waveguide. However, a wide field-of-view that can cover most of the angles is highly desirable. An antenna structure that provides such beam characteristic is proposed in this appendix. A drawing of the antenna is shown in Fig. D.1. The antenna is fed from a dielectric slot waveguide [131], which has the EM field concentrated in the air gap between the two silicon side walls. By flaring the end of the two side walls and connecting two dielectric pucks to the end of the flares, the EM field from the slot waveguide can be diffracted gradually achieving a broadened

APPENDIX D: SLOT OPTICAL ANTENNA
Fig. D.1 Drawing of the proposed wide beam slot optical antenna.

beamwidth. A simulated (FIT) E-field intensity radiated from the antenna at 195THz is shown in Fig. D.2. As seen, the E-field reached to the area in the waveguide direction is relatively small, however, majority of the wafer area is uniformly cover by the radiated field. The simulated (FEM) antenna radiation patterns at 185THz, 195THz and 205THz are shown in Fig. D.3. At 185THz, the E-plane 3dB beamwidth is 180º. As frequency increases, the 3dB beamwidth increased to 240º and 280º at 195THz and 205THz, respectively, which is very close to a real omnidirectional antenna. For the far-field measurement, uniformly separated probe antennas should be placed a distance away from the slot antenna and the transmission performance should be measured to characterize the slot antenna radiation at different angles, as shown in Fig. D.4. The simulated (FIT) transmissions versus different angles (θ) of deviation from the antenna boresight is shown in Fig. D.5. Note that in the simulation, the phase center separation between the slot antenna and the probe antennas is 3µm. As seen, the predicted antenna transmissions show excellent resemblance to the simulated radiation patterns shown in Fig. D.3. Compared to
Fig. D.2 Simulated (FIT) radiated E-field intensity of the proposed slot optical antenna at 195THz.

Fig. D.3 Simulated (FEM) E-plane antenna radiation patterns at 185GHz, 195THz and 205THz.
Fig. D.4 Drawing of the antenna setup used to characterize the slot antenna far-field performance. Note that the probe antennas used are the optical dielectric antennas presented in Chapter 4.

Fig. D.5 Simulated (FIT) transmission performance between the slot antenna and probe antennas at different angle of deviation from the antenna boresight direction.
the measured antenna transmission results shown in chapter 4, due to the smaller antenna gain this simulated performance is about 6dBi lower in boresight direction for the same antenna separation.

To feed this slot antenna, an optical waveguide to slot waveguide transition is designed. Its geometry and E-field distribution are shown in Fig. D.6. This transition introduces about 0.5dB loss throughout the entire 172THz to 222THz bandwidth.

Fig. D.6 Geometry and E-field intensity of the optical waveguide to slot waveguide transition.
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