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UNIVERSITY OF CALIFORNIA, SAN DIEGO

Tunable RF Antennas and Filters for Advanced Communication Systems and Wideband Quasi-Optical Network Analyzer for Millimeter-Wave and Terahertz Applications

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

in

Electrical Engineering (Electronic Circuits and Systems)

by

Abdullah J. Alazemi

Committee in charge:

Professor Gabriel M. Rebeiz, Chair Professor James F. Buckwalter Professor Gert Cauwenberghs Professor Brian G. Keating Professor Daniel F. Sievenpiper

2015

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Chair

University of California, San Diego

2015

DEDICATION

In memory of my mother

With love and eternal appreciation

EPIGRAPH

"Do not follow where the path may lead. Go instead where there is no path and leave a trail."

-Ralph Waldo Emerson

"Yesterday is but today's memory, and tomorrow is today's dream." —Kahlil Gibran

"A father gives his child nothing better than a good education." —Prophet Mohammed (PBUH)

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The dissertation author was the primary author of the work in these chapters, and co-authors (Prof. Gabriel M. Rebeiz, Dr. Hyun-Ho Yang, and Dr. Hong-Ming Lee)

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ABSTRACT OF THE DISSERTATION

Tunable RF Antennas and Filters for Advanced Communication Systems and Wideband Quasi-Optical Network Analyzer for Millimeter-Wave and Terahertz Applications

by

Abdullah J. Alazemi

Doctor of Philosophy in Electrical Engineering (Electronic Circuits and Systems)

University of California, San Diego, 2015

Professor Gabriel M. Rebeiz, Chair

The dissertation comprises of 4 projects; double bow-tie slot antennas for wideband millimeter-wave and terahertz applications, a 100 - 300 GHz free-space scalar network analyzer using compact Tx and Rx modules, a tunable dual-band quadruplepole antenna for carrier aggregation systems, and RF MEMS tunable 4-pole bandpass filters with bandwidth control and improved stopband rejection.

The dissertation first presents millimeter-wave and THz double bow-tie slot antennas on a synthesized elliptical silicon lens. Two different antennas are designed to cover 0.1 - 0.3 THz and 0.2 - 0.6 THz, respectively. The double bow-tie slot antenna results in a wide impedance bandwidth and 78-97% Gaussian coupling efficiency over a 3:1 frequency range. A wideband CPW low-pass filter is designed using slow-wave techniques, and the measured filter response shows an $S_{21} < -25$ dB over a 3:1 frequency range. Absolute gain measurements done at 100 - 300 GHz and 200 - 600GHz confirm the wideband operation of this design. The double bow-tie slot antenna is intended to fill the gap between standard double-slot antennas and log periodic and sinuous antennas, with applications areas in radio astronomy and imaging systems.

In chapter 3, the dissertation presents a 100 - 300 GHz quasi-optical network analyzer using compact transmitter and receiver modules. The transmitter includes a wideband double bow-tie slot antenna and employs a Schottky diode as a frequency harmonic multiplier. The receiver includes a similar antenna, a Schottky diode used as a sub-harmonic mixer, and an LO/IF diplexer. The 100 - 300 GHz RF signals are the 5th to 11th harmonics generated by the frequency multiplier when an 18 - 27 GHz LO signal is applied. The measured transmitter conversion gain with P_{in} = 18 dBm is -35 to -59 dB for the 5th to 11th harmonic, respectively, and results in a transmitter EIRP of +3 dBm to -20 dBm up to 300 GHz. The measured mixer conversion gain is -30 to -47 dB at the 5th to 11th harmonic, respectively. The system has a dynamic range > 60 dB at 200 GHz in a 100 Hz bandwidth for a transmit and receive system based on 12 mm lenses and spaced 60 cm from each other. Frequency-selective surfaces at 150 and 200 GHz are tested by the proposed design and their measured results agree with the simulations. Application areas are low-cost scalar network analyzers for wideband quasi-optical 100 GHz to 1 THz measurements.

In chapter 4, the dissertation presents a tunable dual-feed, dual-band, quadruplepole antenna for carrier aggregation and mobile communication systems. The dual-feed PIFA (planar inverted F-antenna) has four operating frequencies which are independently tunable within the long-term evolution frequency range (LTE) from 0.7 to 2.7 GHz. Tuning is obtained using varactor diodes: two at low-band (0.7 – 1 GHz) and two at high-band (1.6 – 2.6 GHz). The antenna is well matched at both feeds (S₁₁, S₂₂ < -10 dB) and the isolation between the feeds is < -20 dB at low-band and < -13 dB at high-band. The antenna volume is $66 \times 100 \times 3.15$ mm³ and is placed on RO4003C ($\varepsilon_r = 3.55$) printed-circuit board (PCB). The measured radiation efficiency is 25-50% at low-band and 30-62% at high-band with varactor diodes and it can be improved to > 50% using high Q devices such as RF-MEMS (Q > 200). A new concept to achieve tunable wideband performance is demonstrated using an extra shunt variable capacitor.

In chapter 5, 0.7 - 1 GHz and 1.7 - 2.4 GHz tunable 4-pole bandpass filters with bandwidth control and improved stopband rejection using RF-MEMS capacitors and varactor diodes have been demonstrated. The MEMS capacitors are fabricated and fully packaged using a 0.18 μ m CMOS standard process with integrated high voltage drivers and SPI/RFFE control logic and with reliability in the billions of cycles. The 0.7 - 1 GHz filter results in insertion loss < 2.9 dB with controllable 1-dB bandwidth of 30-76% and IIP3 > 25.2 dBm with varactors and \approx 52 dBm with MEMS only. The measured ACPR (adjacent channel power ratio) for a 5-MHz Wideband CDMA signal is at least 49 dB at 10 dBm input power with varactors and at least 58.3 dB at 26 dBm input power with MEMS only. The 1.7 - 2.4 GHz filter results in insertion loss < 3.7dB with controllable 1-dB bandwidth of 70-178% and IIP3 > 20 dBm with varactors and \approx 48 dBm with MEMS only. The measured ACPR for a 5-MHz Wideband CDMA signal is at least 37 dB at 10 dBm input power with varactors and at least 57.8 dB at 26 dBm input power with MEMS only. This paper also discusses the requirements of RF-MEMS capacitors with SPI and RFFE digital control for wireless communication systems.

Chapter 1

Introduction

1.1 Millimeter-Wave and Terahertz Technology

Millimeter-waves frequencies generally correspond to the frequency spectrum between 30-300 GHz, with wavelengths between one to ten millimeters. Sub-millimeter waves (also known as Terahertz) cover the frequency band of 0.3-3 THz (0.1-1 mm wavelength). The electromagnetic radiation at these frequencies has unique and attractive qualities. For example, it can stimulate molecular and electronic motions in many materials leading to its widespread use in spectroscopy and bio-imaging applications. Moreover, mm-wave and terahertz frequencies can yield extremely high resolution images due to their small wave-length. They are also capable of carrying high amounts of data, due to the large absolute bandwidth they are able to offer. They can permit data rates in excess of 10 Gbits/s [1]. These advantages can be extensively used in many different applications such as communication systems, automotive radars and radio astronomy, (1.1).

There are several advantages of using millimeter-waves and terahertz over the microwaves from an antenna design perspective. One of these advantages is the increase



Figure 1.1: Millimeter-wave and terahertz applicatons

of antenna gain (Gain α f^2) which provides narrower pattern beamwidth resulting in high imaging resolution. Another advantage is the small size of components. While integrated circuits (ICs) keep the circuitry small, antennas scale down in size as the frequency of operation increases. Unfortunately, these small wavelengths present new difficulties and challenges in terms of integration, packaging, transition between RF components etc. This will also introduce substantial transmission lines loss and other problems that can be difficult to compensate. Coaxial components become fragile and lossy, and they are often replaced by waveguide components which are bulky, expensive, and inherently band-limited.

Among all these design challenges in millimeter-wave and terahertz frequency, there is an increasing demand for highly-efficient planar antennas which have wide impedance matching and good radiation properties. The potential for power loss into the antenna substrate modes results in poor radiation patterns. To avoid the excessive substrate modes power loss, the thickness of the dielectric must be less than $0.1\lambda_d$ for dipoles and $0.25\lambda_d$ for slots [2]). At millimeter-waves, this will result in thin and fragile substrates which are hard to implement and difficult to manufacture. Another technique was introduced by Rutledge [3] to eliminate the substrate modes, the planar antenna is placed on a dielectric lens. Like the thick substrate layer, a high dielectric constant (ε_r) lens results in nearly unidirectional patterns, and the lens does not support the substrate modes. With dielectric lenses that have high ε_r (> 10), the antenna will radiate most



Figure 1.2: Antenna on different lens positions.

of the power into the lens and only (< 10%) of the power is radiated into the back air. The lens will also provide mechanical rigidity and control the focusing properties of the antenna. In addition, the radiation patterns produced by the dielectric lenses have high Gaussian coupling efficiency (> 75%) which makes such an antenna extremely useful for quasi-optical systems.

There are two important factors that define the performance of lens antennas. The first factor is the feed antenna. For good radiation patterns and efficiency, the feed antenna must radiate in broadside direction, and the E- and H-planes should be symmetrical in the dielectric half-space. For wideband performance, the feed antenna must provide stable input impedance variations with the wide frequency range. The second factor is the dielectric lens shape. The lens can be shaped to control the focusing properties of the antenna.

Several antennas were used on dielectric lenses. Dipoles, slots, and ring antennas [4, 6, 7, 48] produce excellent patterns but they are bandwidth limited (\pm 15%). For wideband solutions, frequency independent and self-complementary antennas such as log-periodic, sinuous, and spiral are used [10, 11, 44, 46]. However, log-periodic and sinuous antennas suffer from high cross-polarization levels (>-15 dB) and spiral antenna only results in circularly polarized patterns.

The dielectric lenses can be hemispherical, hyper-hemispherical, or elliptical based on the attached extension length (Fig. 1.2). The extension length (L_{ext}) at hyper-hemispherical position is approximately $0.28R_{lens}$ while it is $0.33-0.38R_{lens}$ at elliptical

position. When the antenna radiates most of the power into the lens, it was found that the radiation rays are bended by the dielectric lens towards the broadside direction result in sharpening the patterns and effectively increasing the gain of the antenna. The hyperhemispherical or elliptical lenses are capable of coupling well with a Gaussian-beam system. Previous studies proved that any antenna placed on hyper-hemispherical dielectric lens has an of gain by n^2 (*n* is the dielectric index of refraction. On the other hand, any antenna placed at the focus of the elliptical lens will result in far-field patterns with a main-beam that is diffraction limited by the aperture of the lens. At that position, all the radiation waves inside the dielectric will transform from a spherical form to planar waves and the gain of the antenna can be calculated by

$$D = \frac{4\pi\varepsilon_{ap}A_P}{\lambda_o^2} \tag{1.1}$$

where A_P is the lens physical area ($\pi R_{lens}^2/4$) and ε_{ap} is the aperture efficiency.

Nevertheless, wideband feed antennas with broadband patterns and low crosspolarization levels for dielectric lenses remain an important topic of investigation and research.

1.2 Evolution of Mobile Communications

Wireless technology has become an essential part of life in many parts of the world. The exponential increase of traffic in mobile communication networks due to the surge in demand for accessing mobile data in the form of multimedia files and online streaming services on the fly, drove the research to the development and design of next generation mobile communication systems. To satisfy the increasing demand, more bands were added crowding the limited spectrum available. However, increasing the number of bands alone proved to be not enough to support the required operating



Figure 1.3: Carrier aggregation modes.

bandwidth of up to 100 MHz in LTE-Advanced. As a result, new techniques are used in LTE such as carrier aggregation and MIMO systems (Multiple-Input-Multiple-Output) to achieve more bandwidth and therefore higher data rates.

Carrier aggregation is one key enabler of LTE - Advanced to meet the IMT-Advanced requirements in terms of peak data rates of up to 1 Gbps. It is a highly demanded feature from a network operator perspective, since it enables also the aggregation of different spectrum channels. Carrier aggregation has three different modes and meant to achieve a bandwidth of 100 MHz within the LTE frequency range (0.7 - 2.7GHz). The easiest way to arrange aggregation is to use contiguous component carriers within the same operating frequency band, so called intra-band contiguous. For noncontiguous allocation it could either be intra-band, i.e. the component carriers belong to the same operating frequency band, but have a gap or gaps, in between, or it could be inter-band, in which case the component carriers belong to different operating frequency bands (Fig. 1.3).

Carrier aggregation rises a number of technical challenges such as the need for wideband power amplifiers, highly flexible switches and tunable antenna elements to support the many different combinations of channels aggregated over a wide frequency spectrum. To achieve tunability in antennas, they are loaded with one or more tunable devices such as variable capacitors (varactors). The tuners will not only force the antenna to resonate at specific frequency but also tune it during the whole frequency range. Varactors are now extensively used in modern mobile phones and these devices are mostly RF-MEMS varactors or SPNT (single-pole-multiple-throw) switches loaded by a bank of capacitors. In research areas, diodes are used to achieve the same performance. Despite the low Q-factor of diodes (Q < 100) compared to RF-MEMS (Q > 250 at 700 MHz), they have proven adequate to support carrier aggregation in small antennas. The topology of the loaded varactors is a design criteria that needed to be studied further. The varactors can be connected either in series or shunt topologies. This dissertation presents the effects of using series and shunt tuners in small antennas.

1.3 RF-MEMS Technology

Micro-Electro-Mechanical-Systems, or MEMS, are electro-mechanical systems fabricated on a common substrate using micro-fabrication processes. Examples of these systems include sensors, actuators, and RF-MEMS. RF-MEMS are MEMS devices designed to operate at RF to millimeter-wave frequencies (0.1-100 GHz) and commonly used as metal-contact or capacitive-contact switches and variable capacitors. RF-MEMS provide several advantages in terms of size, low cost, and often better performance than their traditional solid-state circuits counterparts. RF-MEMS devices have low-loss, high linearity, high power handling, high isolation, and extremely low power consumption. The switching speed of most electrostatic MEMS switches is 2-40 μ s, and thermal/magnetic switches are 200-3,000 μ s. These advantages allow RF-MEMS to be part of many commercial and military applications such as tunable antennas and filters, phase shifters, reconfigurable matching networks, and satellite communications.

In particular, the variable capacitor, or varactor, is one of the famous RF-MEMS devices and it is commonly used as a lumped component within RF circuits and transceiver platforms. Although the varactors offer high linearity, large tuning range, high Q-factor (low-loss), and virtually zero power consumption, they are fragile and not reliable in medium and long terms. The reliability of mature MEMS switches is 0.1-40 Billion cycles. However, many systems require switches with 20-200 Billion cycles.

Recently, RF-MEMS technology is experiencing a substantial development effort to enhance its capabilities and commercial vendors are now producing MEMS packaged devices with high reliability. This include Cavendish Kinetics and wiSpry for capacitive switches [12, 92], Omron and Radant MEMS for ohmic contact switches [14, 15]. These devices are small ($\approx 2 \text{ mm}^2$ bumped die), hermetically-packaged, and ESD protected with excellent long-term reliability under medium to high RF power conditions (up to 2 W). Cavendish Kinetics switched capacitors, for example, can handle RF power of 33 dBm with cycling > 10B cycles and switching time < 50 μ s. They are also easy to control with SPI or RFFE digital controls.

1.4 Scope of Dissertation

This dissertation presents on details four separate projects in three main research areas, millimeter-wave systems, tunable RF antennas, and tunable RF filters.

Chapter 2 presents millimeter-wave and THz double bow-tie slot antennas on a synthesized elliptical silicon lens. Two different antennas are designed to cover 0.1 - 0.3 THz and 0.2 - 0.6 THz, respectively. The double bow-tie slot antenna results in a wide impedance bandwidth and 78-97% Gaussian coupling efficiency over a 3:1 frequency range. A wideband CPW low-pass filter is designed using slow-wave techniques, and the measured filter response shows an S₂₁ < -25 dB over a 3:1 frequency range. Absolute gain measurements done at 100 - 300 GHz and 200 - 600 GHz confirm the wideband operation of this design. The double bow-tie slot antenna is intended to fill the gap between standard double-slot antennas and log periodic and sinuous antennas, with applications areas in radio astronomy and imaging systems.

Chapter 3 presents a 100 - 300 GHz quasi-optical network analyzer using compact transmitter and receiver modules. The transmitter includes a wideband double bowtie slot antenna and employs a Schottky diode as a frequency harmonic multiplier. The receiver includes a similar antenna, a Schottky diode used as a sub-harmonic mixer, and an LO/IF diplexer. The 100 - 300 GHz RF signals are the 5th to 11^{th} harmonics generated by the frequency multiplier when an 18 - 27 GHz LO signal is applied. The measured transmitter conversion gain with P_{in} = 18 dBm is -35 to -59 dB for the 5th to 11^{th} harmonic, respectively, and results in a transmitter EIRP of +3 dBm to -20 dBm up to 300 GHz. The measured mixer conversion gain is -30 to -47 dB at the 5th to 11^{th} harmonic, respectively. The system has a dynamic range > 60 dB at 200 GHz in a 100 Hz bandwidth for a transmit and receive system based on 12 mm lenses and spaced 60 cm from each other. Frequency-selective surfaces at 150 and 200 GHz are tested by the proposed design and their measured results agree with the simulations. Application areas are low-cost scalar network analyzers for wideband quasi-optical 100 GHz to 1 THz measurements.

Chapter 4 presents a tunable dual-feed, dual-band, quadruple-pole antenna for carrier aggregation and mobile communication systems. The dual-feed PIFA (planar inverted F-antenna) has four operating frequencies which are independently tunable within the long-term evolution frequency range (LTE) from 0.7 to 2.7 GHz. Tuning is obtained using varactor diodes: two at low-band (0.7 – 1 GHz) and two at high-band (1.6 – 2.6 GHz). The antenna is well matched at both feeds (S_{11} , S_{22} < -10 dB) and the isolation between the feeds is < -20 dB at low-band and < -13 dB at high-band.

The antenna volume is $66 \times 100 \times 3.15 \text{ mm}^3$ and is placed on RO4003C ($\varepsilon_r = 3.55$) printed-circuit board (PCB). The measured radiation efficiency is 25-50% at low-band and 30-62% at high-band with varactor diodes and it can be improved to > 50% using high Q devices such as RF-MEMS (Q > 200). A new concept to achieve tunable wideband performance is demonstrated using an extra shunt variable capacitor.

Chapter 5 presents 0.7 - 1 GHz and 1.7 - 2.4 GHz tunable 4-pole bandpass filters with bandwidth control and improved stopband rejection using RF-MEMS capacitors and varactor diodes. The MEMS capacitors are fabricated and fully packaged using a 0.18 μ m CMOS standard process with integrated high voltage drivers and SPI/RFFE control logic and with reliability in the billions of cycles. The 0.7 - 1 GHz filter results in insertion loss < 2.9 dB with controllable 1-dB bandwidth of 30-76% and IIP3 > 25.2 dBm with varactors and \approx 52 dBm with MEMS only. The measured ACPR (adjacent channel power ratio) for a 5-MHz Wideband CDMA signal is at least 49 dB at 10 dBm input power with varactors and at least 58.3 dB at 26 dBm input power with MEMS only. The 1.7 - 2.4 GHz filter results in insertion loss < 3.7 dB with controllable 1-dB bandwidth of 70-178% and IIP3 > 20 dBm with varactors and \approx 48 dBm with MEMS only. The measured ACPR for a 5-MHz Wideband CDMA signal is at least 37 dB at 10 dBm input power with varactors and at least 57.8 dB at 26 dBm input power with MEMS only. This paper also discusses the requirements of RF-MEMS capacitors with SPI and RFFE digital control for wireless communication systems.

Chapter 2

Wideband Double Bow-Tie Slot Antennas for Millimeter-Wave and Terahertz Applications

2.1 Introduction

Double-slot and slot-ring antennas on a quartz and silicon dielectric lenses have been extensively used for planar millimeter-wave and terahertz applications, with applications in radio-astronomy, communication systems and low-power radars [6, 7, 10, 11, 16–30, 44, 46]. They offer symmetrical patterns into the dielectric lens which transfers to a high Gaussian-beam coupling efficiency in a quasi-optical system. Also, they do not support substrate modes and have a low cross-polarization level (< -20 dB) and their geometry allows for a dual-polarization design [16, 17]. The effect of the dielectric lens-to-air interface was also extensively studied versus the extension length [6, 18, 44]. Also, the reflections inside the lens on the far-field patterns and antenna impedance can now be accurately simulated using Ansys-HFSS [31] and are now taken into account in



Figure 2.1: Wideband double bow-tie slot antenna with CPW low-pass filter.

the design procedure of high-performance receivers.

The double-slot antenna naturally focuses the energy to the center of the doubleslot and a detector is placed at this position (Fig. 2.1). A low-pass filter is therefore required at the edge of one of the slot antennas to allow for biasing of the detector and for DC and IF signals. The low-pass filter is typically a multi-section low/high impedance filter which presents a low impedance at the slot feed and allows for a symmetrical field distribution between the two slot antennas, and a symmetrical pattern in the far-field.

One drawback of the standard double-slot antenna is its relatively narrow impedance and pattern bandwidth. The antennas can operate well over a $\pm 20\%$ bandwidth, and while this is acceptable for a large set of applications, it is not acceptable for octave or multi-octave operations. In this case, a self-complementary antenna [21] is used such as log-periodic, spiral or sinuous antennas [7, 10, 11, 22–24, 46]. These antennas can cover a 10:1 frequency range with a near constant impedance on a dielectric lens, and have a natural RF filter between the detector placed at the center and the DC (or low IF) taken from the edges. Also, single and dual polarized designs can be built using such antennas [7, 11, 23, 24]. However, they do not have the polarization purity as double-slot antennas or the ability to control the antenna impedance for a variety of different detectors. Nevertheless, frequency independent, self-complementary antennas are appealing in wideband applications since their bandwidth is determined only by their dimensions. Recently, a double-slot antenna with improved butterfly-shaped design showed a 2:1 response at 0.6 - 1.2 THz [25]. In this design, the width of the slot dipoles is substantially increased to achieve wideband performance. However, the impedance and radiation patterns of this antenna have not been thoroughly analyzed on dielectric lenses.

This chapter presents a double bow-tie slot antenna for wideband applications, with a frequency range of approximately 3:1. The work is intended to fill the gap between the double-slot antenna and the ultra-wideband complementary antennas. The design and measurements of a double bow-tie slot antenna placed on a synthesized elliptical silicon dielectric lens is presented at 100 - 300 GHz and at 200 - 600 GHz with good performance.

2.2 Design and Simulations

The double bow-tie slot antenna on a silicon dielectric lens is presented in Fig. 2.1. There are three essential design criteria which are required for wideband operation and these are: a) wideband impedance at the detector ports, b) a low-pass filter capable of operating at mm-wave frequencies and presenting a low impedance to the slot-antenna feed point over a 3:1 frequency range, and c) wideband and symmetrical patterns with low cross-polarization levels and capable of a high Gaussian-beam coupling efficiency. The wideband low-pass filter rejects any RF power leaking out of the double bow-tie slot antenna and increases the antenna efficiency.

2.2.1 Impedance

The double bow-tie slot antenna geometry is first optimized assuming a perfect short at both antennas as shown in Fig. 2.2. Several bow-tie angles were considered starting from $\phi = 0^{\circ}$ (standard double-slot) to $\phi = 45^{\circ}$. Note that the maximum allowed



Figure 2.2: (a) 100 - 300 GHz Double bow-tie slot antenna layout, (b) design schematic, (c) antenna return loss vs. different bow-tie angles (ϕ) with $Z_{in} = 100 \Omega$ and S = 0.

angle before the edges touch each other is $\phi \approx 55^{\circ}$. The separation between the bow-tie antennas is $0.17\lambda_o$ at 200 GHz (100 – 300 GHz operation) which is a compromise between pattern degradation versus frequency and antenna impedance. A larger separation allows for wider bow-tie angles and less impedance variation, but results in rapid pattern degradation at the higher edge of operating band. A smaller separation has higher mutual coupling between the two slot antennas and also smaller bow-tie angles, and this results in larger antenna impedance variation.

The double bow-tie slot antenna is designed at a center frequency (f_o) of 200 GHz with L = 0.56 mm, S = 0.26 mm, and $\phi = 42^\circ$. The 200 – 600 GHz antenna is designed for $f_o = 400$ GHz with L = 0.28 mm and S = 0.136 mm. The simulated input impedance of different bow-tie antennas vs. frequency is shown in Fig. 2.2(c). Simulations are done using ADS-Momentum [32] on a dielectric half-space with ($\varepsilon_r =$


Figure 2.3: S_{11} of 200 – 600 GHz antenna with $Z_{in} = 100 \Omega$.

11.7), and do not take into account any lens reflections. The bow-tie slot antennas operate near their second resonance point due to their inherent wideband operation at this length [29]. Note that slot antennas with an angle of $\phi = 20 - 45^{\circ}$ are much more wideband than the standard $\phi = 0^{\circ}$ design. The mutual coupling (Z_{12}) is taken into account in the driving point impedance of each antenna as $Z_{ant} = Z_{11} - Z_{12}$ [28], and a transmission-line section of $Z_o = 60 \Omega$ is used between the antennas and the detector for wideband matching. The impedance match is < -10 dB at 120 - 350 GHz for a differential load of 100 Ω . In this design, the detector impedance is chosen to be 100 Ω , but other detector impedances can be used and the matching network re-optimized. The input return loss of 200 - 600 GHz antenna is shown in Fig. 2.3. Due to difficulty of measuring the input impedance within the wideband frequency range, a low frequency double bow-tie slot antenna is placed on RO3010/6010 printed circuit board to cover a wideband frequency range of 2 - 14 GHz.

2.2.2 Wideband Low-Pass Filter

The low-pass filter is an essential component in the design and must have high rejection over the full operating band since it should present a short to the slot-antenna feed. The standard high-Z/low-Z impedance filter which is commonly used is narrow-



Figure 2.4: (a) CPW low-pass filter layout (gap is white, air-bridges are not shown), (b) zoom-in of the filter, (c) simulated S-parameters of the wideband low-pass filter with input impedance = 50Ω and Smith chart for 115 - 285 GHz.

band with a maximum rejection of 20 - 25 dB and a second pass-band at $\approx 3f_0$, where f_0 is the 3-dB cutoff frequency of the filter. Therefore, a complex miniature multisection coplanar-waveguide periodic filter has been designed and simulated using ADS-Momentum. The filter is based on a slow-wave propagation design which is accomplished by effectively increasing the transmission-line *L* and *C* values as shown in Fig. 2.4 [33]. This is achieved by changing the CPW line width to control the inductance per unit length and changing the space between the signal lines and the ground to control the capacitance. The overall footprint of the periodic structures remains the same compared to a standard 50 Ω CPW line. Three CPW unit cells are designed to reject signals at 100 - 300 GHz, and each unit-cell length is determined by its cutoff frequency f_c with

$$l_{1,2,3} = \frac{c_0 / \sqrt{\varepsilon_{eff}}}{2f_{c_{1,2,3}}} \quad (\varepsilon_{eff} \approx 6.35)$$
(2.1)

The length of each unit cell is optimized such that $l_1 = 0.27$ mm, $l_2 = 0.25$ mm, and $l_3 = 0.22$ mm and f_{c_1} , f_{c_2} , $f_{c_3} = 220$, 238, 270 GHz, respectively. The simulated S-parameters show a rejection > 20 dB at 110 GHz - 320 GHz with an insertion loss < 0.5 dB up to 50 GHz (Fig. 2.4 (c)). The filter is 0.78 mm long with line widths and slot widths varying between 3.5 μ m to 18.5 μ m and is compatible with standard lithography. The impedance seen from the bow-tie slot antenna is < 1-j50 Ω at 115 - 280 GHz (Fig. 2.4 (c)). The RF low-pass filter for the 300 - 600 GHz antenna has nearly the same performance (S₂₁ < -20 dB). These filters exhibit low insertion loss and simple fabrication.

2.2.3 Radiation Patterns

The simulated patterns inside the dielectric lens for the double bow-tie antenna with $L = 0.37\lambda_o$ and $S = 0.17\lambda_o$ at 200 GHz are shown in Fig. 2.5. Again, these



Figure 2.5: Simulated radiation patterns inside the silicon at: (a) 125 GHz, (b) 175 GHz, (c) 225 GHz. (d) 275 GHz.

patterns are simulated on a semi-infinite substrate using ADS-Momentum [28–30]. The expected unsymmetrical patterns due to the filter radiation are minor of 100 - 300 GHz. The cross-polarization component is < -40 dB in the principal planes and remains < -20 dB at 100 - 300 GHz in the diagonal planes. The power radiated to the backside is < 6% over the frequency range. These patterns are similar to double-slot antennas with $\phi = 0^{\circ}$ (standard design).

A better way to look at the pattern is to study the far-field patterns when the double bow-tie slot antenna is placed at the hyper-hemispherical point with $L_{ext} = R_{lens}/n$ (Fig. 2.6), where L_{ext} is the extended length, R_{lens} is the lens radius, and n is the silicon index of refraction (≈ 3.42). At the hyper-hemispherical position, the directivity increases by n^2 and at the synthesized elliptical position ($L_{ext} = 0.33-0.38R_{lens}$), the antenna becomes diffraction limited with



Figure 2.6: (a) Simulated double bow-tie slot antenna directivity at different lens positions, (b) aperture and Gaussian coupling efficiencies for double bow-tie slot antenna and standard double slot antenna.

$$D = \frac{4\pi\varepsilon_{ap}A_P}{\lambda_o^2} \tag{2.2}$$

where A_P is the lens physical area ($\pi R_{lens}^2/4$) and ε_{ap} is the aperture efficiency.

For a lens radius of 6 mm, the elliptical position is achieved with $L_{ext} = 0.34R_{lens}$ and the far-field patterns are simulated using Ansys-HFSS. A comparison between the antenna directivity at different lens positions is shown in Fig. 2.6(a) and the aperture efficiency and Gaussian coupling efficiency at the synthesized elliptical position are presented in Fig. 2.6(b). The Gaussian beam is done using

$$|E_{Gaussian}(\theta)| = e^{-(\theta/\theta_o)^2}$$
(2.3)

The Gaussian coupling efficiency and aperture efficiency are 78-97% and 55-87% at 100 - 300 GHz, respectively. The degradation in aperture efficiency at 300 GHz is due to the directive and non-symmetrical patterns inside the silicon lens. Note that the double bow-tie slot antenna results in very similar values as the standard double slot antenna.

2.3 Measurements

2.3.1 Impedance

An scaled double bow-tie slot antenna model with $f_o = 8$ GHz was fabricated on a 20 mils thick Rogers 3010/6010 substrate with L = 40 mm and S = 2.5 mm, and placed on a 10 cm silicon lens. The low-pass filter was not present and a coax-feed was soldered to the two antenna feeds (Fig. 2.7). A short-open calibration was done at the coaxial line tips, and the measured impedance is referenced to the antenna ports. Also, time domain was used to remove the multiple reflections from the silicon-air interface. The two coax lines were fed into a differential vector network analyzer to measure S₁₁. The measured input impedance is equivalent to $2 \times (Z_{11} - Z_{12})$ and agrees well with simulations and shows a wideband operation from 2 - 14 GHz.

2.3.2 Wideband CPW Low-Pass Filter

The CPW low-pass filter was fabricated on a high-resistivity silicon substrate ($\varepsilon_r = 11.7$, $\rho > 7 \ k\Omega - cm$) with 0.2 μ m SiO₂ layer. A sputtered gold layer (0.3 μ m) with additional electroplated gold (2 - 3 μ m) on some areas was used. Air-bridges with $l = 125 \ \mu$ m and $w = 20 \ \mu$ m were placed every 80 - 120 μ m over the filter to equalize the CPW grounds, and their capacitance was taken in the filter design. The measured filter



Figure 2.7: Measured and simulated S_{11} with differential input impedance $Z_{in} = 100 \Omega$.



Figure 2.8: Measured S-parameters of the wideband CPW low-pass filter.

response shows a wideband rejection ≥ 20 dB at 100 – 220 GHz (max. range), but the insertion loss is higher than simulated and is 1.5 dB at 10 GHz. We believe that this is due to charges created at the SiO₂/silicon interface in the CPW gaps, while greatly increase the transmission-line loss (see [34] for more details). The filter is well matched with S₁₁ < -20 dB. Since the filter is symmetrical, S₂₂ and S₁₂ are not shown for brevity.

2.3.3 Radiation Patterns

Two double bow-tie slot antennas were fabricated on a high-resistivity silicon substrate using sputtered and electroplated gold layers. One design was centered at



Figure 2.9: Double bow-tie slot antenna measurement setup.

Frequency (GHz)	Horn Antenna Gain (dB)	AMC-VDI No.
102-130	24	332 (WR-10-6)
123-160	24	333 (WR-6)
170-215	21	335 (WR-5)
220-265	26	332 (WR-2.8)
246-314	26	333 (WR-2.8)
312-390	26	332 (WR-2.2)
370-470	26	333 (WR2.2)
440-525	26	334 (WR-2.2)

 Table 2.1: Measurement setup



Figure 2.10: Simulated (a,b,c,f) and measured (d,e,f,i) radiation patterns on synthesized elliptical silicon lens with $R_{lens} = 6$ mm. (Solid: E-plane, Dashed: H-plane)

200 GHz (100 – 300 GHz) and another design was centered at 400 GHz (200 – 600 GHz). A nickel/chrome micro-bolometer detector with a size of 3 μ m (width) × 30 μ m (length) × 25 nm (thickness), an impedance of 60 to 120 Ω (different fabrication lots) and a responsivity of 15-35 V/W was used. To avoid the thermal loss to the substrate and maximize the responsivity, an air-suspended bolometer structure was used. Detailed fabrication information and measurement results for this detector are presented in [35]. The 100 – 300 GHz antenna was placed on R_{lens} = 6 mm silicon lens and the 200 – 600 GHz was placed on R_{lens} = 3 mm silicon lens at the synthesized elliptical position (L_{ext} = 0.34R_{lens}), and no matching layer is used. The measurement setup is shown in Fig. 2.9. A selection of horn antennas and AMC-VDI multiplier chains are used to cover 100 – 500 GHz as shown in Table. 2.1. The distance (R_{FF}) between the two antennas is always larger than the far-field condition (2D²/ λ) where D is the diameter of the largest antenna. For the 100 – 300 GHz design, R_{FF} is chosen to be 30 cm. For the 200 – 600



Figure 2.11: Measured and simulated 3-dB beamwidths for the E and H planes ($R_{lens} = 6 \text{ mm}$).



Figure 2.12: Comparison between measured patterns at 225 GHz and a Gaussian beam ($R_{lens} = 6 \text{ mm}$).

GHz design, R_{FF} is 18 cm.

The measured and simulated radiation patterns are shown in Fig. 2.10 for the 100 - 300 GHz design. Due to the mount, the pattern could only be measured to $\pm 25^{\circ}$, but this was enough to check the pattern shape and symmetry. At 125 GHz, the patterns are relatively broad due to the small lens size $(5\lambda_o)$, and at 275 GHz, the patterns start becoming unsymmetrical due to the feed pattern inside the lens. The 3-dB beamwidth for the E and H planes are shown in Fig. 2.11 and the measurements agree well with simulations. A comparison between the measured patterns at 225 GHz and a filled Gaussian beam is shown in Fig. 2.12, with Gaussian coupling efficiency > 97%.

The directivity for the 100 - 300 GHz antenna is shown in Fig. 2.13. Since the



Figure 2.13: Measured and simulated 100 - 300 GHz antenna directivity and gain ($R_{lens} = 6$ mm).



Figure 2.14: Measured radiation patterns (a,b,c) for the 200 - 600 GHz on synthesized elliptical silicon lens with $R_{lens} = 3$ mm. (Solid: E-plane, Dashed: H-plane), (d) microscopic image of the antenna. (units in μ m)

patterns are measured only in two principle planes ($\phi = 0^{\circ}, 90^{\circ}$), a comparison between the simulated and measured directivity is done by extracting it from both simulations and measurements with [36]

$$D = \frac{32400}{\phi_{3dB-E}\phi_{3dB-H}}$$
(2.4)

The measured directivity is 23 - 30 dB at 125 - 275 GHz. The patterns for the 200 - 600 GHz antenna are shown in Fig. 2.14 and show similar directivity as the 100 - 300 GHz antenna (Fig. 2.15). For both antennas, the cross-polarization component could not be measured in the principal planes and was < -30 dB at all frequencies.

2.3.4 Absolute Gain Measurements

A direct way to measure the wideband properties of the double bow-tie slot antenna is to measure its absolute gain vs. frequency. The power received at the detector (P_R) is calculated using the Friis transmission formula [36] as:

$$P_R = P_T G_T G_R (\frac{\lambda}{4\pi R})^2 \tag{2.5}$$

where P_T and G_T are the transmit power and transmit horn gain and are calibrated independently. The double bow-tie slot antenna gain, G_R , is calculated as:

$$G_R = \varepsilon_{br} \varepsilon_{sloss} \varepsilon_{refl} (1 - |\Gamma|)^2 D_R$$
(2.6)

where D_R is the front-pattern directivity obtained using HFSS, ε_{br} is the backside power loss (0.3-0.4 dB at 100 – 300 GHz), ε_{sloss} is the silicon lens absorption loss (0.2 dB), ε_{refl} is the silicon-to-air reflection loss (≈ 1.7 dB) [6], and Γ is the antennadetector impedance mismatch.



Figure 2.15: Measured and simulated 200 - 600 GHz antenna directivity and gain ($R_{lens} = 3$ mm).

The measurement setup is shown in Fig. 2.16 where the source is square-wave AM modulated at 1 kHz which is well within the micro-bolometer detector thermal frequency response. The detector has a measured responsivity of 18.5 V/W and 31.5 V/W for 200 and 400 GHz designs, respectively. Measurements are done at 1 mA of bias current, and the detected voltage is measured using a lock-in amplifier. At 100 – 500 GHz, $P_R = 0.5$ -1 mW, $G_T = 21$ -26 dB, $R_{FF} = 20$ -35 cm depends on the size of the lens, and the measured voltage from the detector is 40 – 80 μ V corresponding to 1.26 – 4.32 μ W of received power.

The receiver antenna gain (G_R) can be accurately obtained using the detector voltage and the above equations, and is plotted in Figs. 2.13 and 2.15. As can be seen, G_R agrees well with simulations over a wide frequency range, showing that this antenna is wideband in pattern and impedance. The ripples in the measured gain are due to the absence of a wideband matching layer, and standing waves in the measurement system.

2.4 Conclusion

Two double bow-tie slot antennas are shown to cover a range frequency range of 100 - 300 and 200 - 600 GHz for millimeter-wave and terahertz applications. The use



Figure 2.16: Measurement setup for wideband double bow-tie slot antennas.

of a bow-tie slot significantly increases the impedance bandwidth, and a new design for the low-pass filter ensures a wideband short over a 3:1 frequency range. This antenna should find applications in radio-astronomical systems or THz imaging systems required a 3:1 frequency range.

2.5 Acknowledgment

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Chapter 3

A 100 – 300 GHz Free-Space Scalar Network Analyzer Using Compact Tx and Rx Modules

3.1 Introduction

Millimeter-wave network analyzers above 110 GHz have been limited to standard waveguide frequency bands since high performance multipliers, mixers and couplers at > 110 GHz are built using waveguide Schottky diodes and waveguide assemblies [37–42]. This is in contrast to coaxial-based instruments where 1 mm coaxial lines allow for DC-110 GHz network analyzers mostly due to their single-mode operation over a wide frequency range. The banded operation above 110 GHz, while acceptable to most devices-under-test, still poses a challenge when wideband operation is required, such as measuring the dielectric constant and loss of materials at 100 – 1000 GHz, or the response of frequency selective surfaces (FSS), quarter-wave plates, and polarizers, for example. These quasi-optical components are especially well suited to a free-space



Figure 3.1: Quasi-optical network analyzer architecture for 100 – 300 GHz systems.

network analyzer since their frequency response is given by plane-wave inputs and outputs, and not by transmission-line port definitions.

There has been several attempts to build wideband network analyzers above 100 GHz. Network analyzers based on non-linear transmission lines (NLTLs) and sampling circuits as picosecond pulse generators and detectors are demonstrated [37, 38]. Different network analyzers have been also demonstrated at millimeter-wave frequencies using optoelectronic techniques and short optical pulses, external electro-optic sampling and heterodyne mixing with small linewidth lasers [39–41]. Recently, Grichener et al. demonstrated a 15 – 50 GHz scalar free-space network analyzer using sinuous antennas and harmonic mixing [42]. The design uses high-order harmonic mixing to transmit and receive multiple RF frequencies at the same time where the RF frequencies are the prime harmonic multiples of the LO frequency and are each down-converted to independent IF channels in the receiver.

Millimeter-wave and terahertz sources and receivers are commonly built using planar antennas and Schottky diodes. This technology has significant advantages in terms of output power, bandwidth and efficiency [43–50]. They are also smaller, lighter,



Figure 3.2: (a) Multiplier RF harmonics from an 18 – 27 GHz LO signal, (b) mixer between RF and LO with (ΔF_{IF} = 30 MHz).

and less expensive than waveguide components. They offer practical solutions for higher frequencies (0.1 - 3 THz) and can be arrayed in linear or two-dimensional arrays without a large increase in weight and cost of the system. Planar antennas such as double-slots, log-periodic, spiral, and sinuous on dielectric lenses have been successfully used for planar receivers at millimeter-wave and terahertz frequencies [6, 18, 20, 44–51]. These antennas exhibit high Gaussian coupling efficiency (> 80%) and do not support substrate modes. This paper extends on the work of [42] with an optimized design for 100 - 300 GHz operation. In this design, a double bow-tie slot antenna is used on a dielectric lens for operation over a 3:1 frequency band. Also, compact and improved harmonic mixers and multipliers are built for the 100 - 300 GHz range. The system allows for a dynamic range of > 60 dB in a 100 Hz bandwidth with a silicon lens radius of 6 mm, a range of 60 cm, and without the use of large collimating Gaussian beam lenses. Wideband measurements on frequency selective surfaces (FSS) are done with accurate results.

3.2 Design

3.2.1 System Architecture

The free-space scalar network analyzer is based on harmonic multipliers for the transmitter and harmonic mixing at the receiver (Fig. 3.1). For the transmitter, a Schottky diode is used at the feed point of a double bow-tie slot antenna with a local oscillator at 18 - 27 GHz. The local oscillator power is 10 - 20 dBm, and is large enough to generate a whole set of harmonics simultaneously starting from the 2nd to the 11th harmonics and higher. Of particular interests are the 5th (90 - 135 GHz), 7th (126 - 189 GHz), 9th (162 - 243 GHz) and 11th (198 - 297 GHz) harmonics, which can cover the entire 90 - 300 GHz range. The receiver also uses harmonic mixing, but in this case, a diplexer is attached to the antenna port so as to separate the local oscillator from the resulting IF signal (DC - 330 MHz). The receiver local oscillator is chosen to be off-set from the transmitter local oscillator, and this offset frequency (ΔF_{IF}) determines the system harmonic number which is used ($5\Delta F_{IF}$, $7\Delta F_{IF}$, etc.). The transmitter and receiver components were fabricated and integrated together on single silicon substrates result in a planar monolithic architecture.

Note that the transmitter radiates all the harmonic components simultaneously as shown in Fig. 3.2(a) and that the receiver detects all the harmonics components at the same time. Therefore, in order to separate the harmonic products using a simple $N \times \Delta F_{IF}$ method and to avoid the mixing spurs landing on the top of the desired IF signal, only the prime radiating harmonics are used (5,7,11,13, etc.). Fig. 3.2(b) presents the resulting mixing spurs for a 5th harmonic transmitter and receiver. Note that none of the mixing spurs lands at $5\Delta F_{IF}$. An exception is done for the 9th harmonic since some different mixing spurs at the third harmonic may land on the desired IF frequency of 9LO_{RX}-9LO_{TX}. However, the transmitter and receiver poorly radiate and receive the 3rd harmonic signal since the antennas are much less efficient and mismatched at this relatively low frequency, and therefore, this does not result in error in the system response.

3.2.2 Antenna Design and Measurements

A double bow-tie slot antenna on a silicon lens is used together a multi-step transformer and results in a wideband 50 Ω impedance (S₁₁ < -10 dB) at 130 – 300 GHz (Fig. 3.3). While a 50 Ω impedance is not necessarily ideal for fundamental mmwave Schottky diode multipliers and mixers, it is found that a near-constant impedance is preferred for wideband harmonic mixers since this is not a tuned approach (see Section C). The double bow-tie slot antenna is attached to a 3-section CPW low-pass filter with a cutoff frequency of 90 GHz which results in a rejection > 20 dB at 100 – 350 GHz (Fig. 3.5) and a loss of 2-3 dB at 18 – 27 GHz. This ensures that the harmonic power generated by the diode does not leak from the LO port and is therefore radiated by the antenna. Detailed information and measurements for the antenna and filter are presented in [52].

The double-slot antenna is placed on a silicon lens with a radius of 6 mm at the synthesized elliptical position ($L_{ext} = 0.34R_{lens}$) [6], and no matching is used at the silicon-air interface. This position is chosen since it couples most efficiently to a plane-wave and allows for a free-space measurement system without the use of external lenses. The antenna patterns and gain versus frequency are measured at 100 – 300 GHz using a 60 Ω micro-bolometer detector, and they include the silicon reflection loss and mismatch between the antenna impedance and the detector (3.4). The measured E and H-plane radiation patterns at 200 GHz are shown in Fig. 3.4(a) with E and H planes with a Gaussian coupling efficiency > 90%. The measured and simulated directivity and gain at 100 – 300 GHz are shown in Fig. 3.4(b) with good agreement. The directivity



Figure 3.3: (a) Double bow-tie antenna with multi-step transformer and Schottky diode, (b) antenna return loss (S_{11}) seen by the Schottky diode with $Z_{in} = 50 \Omega$.

is obtained by [36]

$$D = \frac{32400}{\phi_{3dB-E}\phi_{3dB-H}} \tag{3.1}$$

The antenna gain is important since it determines the transmitter EIRP (P_TG_T) and the power received at the receiver port. Note that the antenna gain at the 3rd harmonic frequencies ($f_{LO} = 50 - 80$ GHz, G = 13-16 dB) is much lower than the 24-26 dB gain at 200 - 300 GHz due to antenna impedance mismatch and the relatively small antenna aperture in terms of wavelengths.

3.2.3 Harmonic Multiplier Design

A Teledyne Scientific millimeter-wave single-ended Schottky diode is used with parameters shown in Table 3.1 and with a cutoff frequency, f_c , of



Figure 3.4: (a) Measured antenna radiation patterns at 200 GHz ($R_{lens} = 6 \text{ mm}$), (b) measured and simulated antenna directivity and gain versus frequency.



Figure 3.5: (a) Transmitter microphotograph with double bow-tie slot antenna, CPW low-pass filter, and Schottky diode, (b) circuit model and Thevenin equivalent-circuit of the transmitter. The receiver is similar to the transmitter except for an LO/IF diplexer.

$$f_c = \frac{1}{2\pi R_s (C_{j0} + C_p)} \approx 670 \text{ GHz}$$
 (3.2)

The diode is attached to the center of the double bow-tie slot antenna using silver epoxy (Fig. 3.5(a)). A simulation environment which includes the antenna impedance, the S-parameters of the CPW low-pass filter, and the diode DC bias at 0.5-0.9 V is used in Agilent-ADS [32] as shown in Fig. 3.5(b). The simulated multiplier conversion gain versus the LO power with different harmonics is shown in Fig. 3.6. A conversion gain of -33 to -55 dB is obtained at the 5th to 11th harmonic for an LO power of 18 dBm (Fig. 3.7). The LO power in simulation is defined as a 50 Ω port and does not include the CPW low-pass filter loss (2-3 dB at 18 – 27 GHz). The conversion gain of the harmonic multiplier is simulated by harmonic balance in Agilent-ADS and calculated using,



Table 3.1: TSC-SS-0112 Schottky diode parameters

Figure 3.6: Simulated multiplier conversion gain versus LO power.

$$P_{LO,avail} = \frac{|V_{LO}|^2}{8Rel\{Z_e\}}$$
(3.3)

$$P_{RF}(n) = \frac{|V_{RF}(n)|^2}{2} Rel\{\frac{1}{Z_e}\}$$
(3.4)

$$CG_{MULT}(n) = \frac{P_{RF}(n)}{P_{LO,avail}}$$
(3.5)

where Z_e is a combination of the CPW low-pass filter and antenna impedance, and is nearly 50 Ω at the LO frequency of 18 – 27 GHz, and also close to 50 Ω at the 5th to 11th harmonic due to the wideband nature of the double bow-tie slot antenna.

3.2.4 Harmonic Mixer Design

The harmonic mixer requires an external diplexer to separate the LO from the IF frequency and this is achieved using a low-pass/high-pass design in a CPW configuration



Figure 3.7: Simulated multiplier and mixer conversion gain versus frequency with LO of 18 - 27 GHz and $P_{LO} = 18$ dBm.

(Fig. 3.8). The IF bandwidth is 3 GHz, but in reality, an IF of 50 - 350 MHz is used. The diplexer is fabricated on a high resistivity silicon substrate using a 3 μ m gold electroplated layer and have a measured loss of 3-4 dB at 30 GHz (Fig. 3.9). The isolation between the IF and LO ports is < 35 dB in the IF and LO frequency range.

The same Teledyne Schottky diode is used and the harmonic mixer is simulated in the impedance environment described above. The simulated mixer conversion gain for different harmonics is shown in Fig. 3.10. A conversion gain of -26 to -43 dB is obtained at the 5th to 11th harmonic using an LO power of 18 dBm. The LO power in simulation is defined as a 50 Ω port and does not include the CPW low-pass filter and diplexer losses. The conversion gain of the harmonic mixer is simulated by harmonic balance in Agilent-ADS and calculated using,

$$P_{RF,avail} = \frac{|V_{ANT}|^2}{8Rel\{Z_e\}}$$
(3.6)

$$P_{IF}(m,-1) = \frac{|V_{IF}(m,-1)|^2}{2} Rel\left\{\frac{1}{Z_e}\right\}$$
(3.7)

$$CG_{MIX}(m,-1) = \frac{P_{IF}(m,-1)}{P_{RF,avail}}$$
(3.8)



Figure 3.8: (a) Diplexer layout on a silicon substrate, (b) zoom-in into the diplexer high-pass and low-pass filters. (Air-bridges are not shown).

where Z_e is a function of frequency, and matched to 50 Ω in the LO/IF band and close to 50 Ω in the RF band.

3.3 System-Level Measurements

The transmitter and receiver are fabricated using 3 μ m electroplated gold on a high-resistivity silicon substrate ($\varepsilon_r = 11.7$, $\rho > 7 k\Omega - cm$) with 0.2 μ m SiO₂ layer. Airbridges with $l = 125 \mu$ m and $w = 20 \mu$ m were placed every 80 - 120 μ m over the CPW filters and at the antenna feed points to equalize the CPW grounds, and their capacitance was taken in the design. A 0.16 mm wide opening is used at the center of the double bow-tie slot antennas and this allows the diodes to be attached using silver epoxy (Fig. 3.5).

Each silicon chip is attached to a 3×4 printed-circuit board (PCB) using 20



Figure 3.9: Simulated LO/IF diplexer S-parameters.



Figure 3.10: Simulated mixer conversion gain versus LO power.



Figure 3.11: Measurement setup for: (a) transmitter and horn antenna as a receiver, (b) receiver and horn antenna as a transmitter, (c) transmitter and receiver together, (d) measurement setup photograph of (b), (e) a photograph of the silicon lens attached to the receiver.

mils RO4003C ($\varepsilon_r = 3.55$, $tan\delta = 0.0027$ at 2.5 GHz). Three wirebonds are used to connect the CPW transmission line (G-S-G) on the silicon chip to the microstrip lines on the PCB and twenty wirebonds are connecting the grounds together. Note that the PCB ground is preventing any radiation from the transmission lines to interfere with the antenna radiation. An SMA connector is used to connect the PCB transmission line to a coaxial cable.

3.3.1 Transmitter

The transmitter chip is shown in Fig. 3.5 and is placed on a silicon lens using a specially designed mount with an LO feed (18 - 27 GHz). The measurement setup of the transmitter is shown in Fig. 3.11(a). The diode DC bias is applied using a bias-T (5542-230 Picosecond-pulse labs) at 0.7-0.9 V. A set of horn antennas operating at 80



Figure 3.12: (a) Measured transmitter ERIP versus frequency with LO of 18 - 27 GHz and $P_{LO} = 18$ dBm, (b) measured multiplier conversion gain versus LO power with $f_{LO} = 23$ GHz.

-300 GHz is used to receive the harmonics radiated from the transmitter and VDI subharmonic mixers are employed to down-convert the radiated LO harmonics. Since the horn antenna gains and the sub-harmonic mixer loss are known, the transmitter EIRP can be measured for different harmonic products by using

$$EIRP = P_T G_T = \frac{P_R}{G_R} (\frac{4\pi R}{\lambda_0})^2$$
(3.9)

and it includes the antenna-to-diode impedance mismatch at the harmonic frequencies, the antenna aperture efficiency, the silicon-to-air matching loss, the back-side power loss (0.3-0.4 dB at 100 - 300 GHz), and the silicon lens absorption loss (0.2 dB). A maximum EIRP of 3 dBm is obtained at 120 GHz and drops to -19 dBm at 290 GHz (Fig. 3.12(a)). The measured values are within \pm 3 dB of the simulated values over the entire frequency range (not shown).

Knowing that $EIRP = P_TG_T$ and since the gain of the double bow-tie slot antenna on a dielectric lens has already been measured (Fig. 3.4(b)), the transmitted power is found and the conversion gain of the transmitter multiplier can be extracted using (3.3–3.5). The conversion gain at the 5th to the 11th harmonic versus the input LO power are shown in Fig. 3.12(b). The input LO power is defined at the connector and includes the CPW low-pass filter loss as in Fig. 3.5. The results show a conversion gain of -35 to -59 dB for the 5th to 11th harmonic for an LO power of 18 dBm at $f_{LO} = 23$ GHz (Fig. 3.12(b)).

3.3.2 Receiver

The receiver chip is also placed on a silicon lens with an LO feed and an IF output, and its measurement setup is shown in Fig. 3.11(b). A selection of horn antennas at 70 - 300 GHz is used to transmit the RF signals using VDI amplifier/multiplier chains



Figure 3.13: Measured mixer conversion gain versus LO power with $f_{LO} = 23$ GHz.



Figure 3.14: Measured down-converted IF power as a function of available RF power with LO of 18 - 27 GHz and $P_{LO} = 18$ dBm.

(AMCs: 332-335). A DC voltage of 0.7-0.9 V is applied to bias the diode using a bias-T with the IF signal path. The conversion gain is measured by first determining the available power over the entire lens aperture ($R_{lens} = 6 \text{ mm}$) by

$$P_{RF,avail} = S \times \pi R_{\rm lens}^2 \tag{3.10}$$

where *S* is the incident plane wave power density $(P_tG_t/4\pi R^2)$, and then measuring the IF power at the output of the receiver. The conversion gain is defined as $(P_{IF}/P_{RF,avail})$ and it includes the same loss components described in the transmitter, and the IF filter loss at 100 – 330 MHz (measured to be 1 dB due to the thin metal layer used). A conversion gain of -30 to -47 dB is obtained for the 5th to 11th harmonic for an LO power of 18 dBm at $f_{LO} = 23$ GHz (Fig. 3.13). Again, the input LO power is defined at the connector and includes the CPW low-pass filter and diplexer losses and the measurements agree well with simulations. The measured down-converted IF power as a function of available RF power for different harmonics with LO of 18 – 27 GHz and $P_{LO} = 18$ dBm is also shown in Fig. 3.14.

The measured conversion gain of the transmitter and receiver versus frequency is shown in Fig. 3.15. The LO frequency is swept from 18 - 27 GHz to cover the whole 100 - 300 GHz range. An LO power of 18 dBm is used for both systems. This measured conversion gain is defined at the input of the connector and includes all the system losses and therefore it is within 3-5 dB lower than simulations (no losses) in Fig. 3.7. It is shown that the conversion gain for every harmonic is almost constant through the LO frequency range.



Figure 3.15: Measured multiplier and mixer conversion gain versus frequency with LO of 18 - 27 GHz and $P_{LO} = 18$ dBm.

3.3.3 System Dynamic Range and Free-Space Measurements

The transmitter and receiver are placed 60 cm from each other, and an absorber screen is used with a 4 × 4 cm hole for placement of the device under test (FSS, etc.) as shown in Fig. 3.11(c). An LO power of 18 dBm is used at both the transmitter and receiver with IF of 30 MHz for 100 – 300 GHz operation (5th – 11th). Screenshots from the spectrum analyzer are presented in Fig. 3.16. A measured signal to noise ratio of 40-80 dB was obtained at 100 – 300 GHz in a 100 Hz bandwidth. The phase noise close to the 5th harmonic carrier is \approx -72 dBc/Hz at 1 kHz offset using an Agilent E8257D synthesizer with the ultra-low noise option. The received IF power levels for 6th and 8th harmonics are lower than the prime harmonics and this is due to the different mixing spurs landing on the top of the desired IF signal and this is why they are not used (Fig. 3.16(b)).

3.3.4 Frequency Selective Surfaces (FSS)

In order to test the performance of the 100 - 300 GHz quasi-optical scalar network analyzer, an FSS is designed and fabricated to be used as a DUT. The FSS sheets consist of a n \times n array of unit cells as shown in Fig. 3.17. Two dual-polarized half-



Figure 3.16: Measured down-converted RF harmonics on spectrum analyzer: (a) S/N ratio at 5th harmonic. (b) RF harmonics in the IF band.



Figure 3.17: A layout of a second order frequency selective surfaces on quartz.

F_0 (GHz)	L (mm)	S (mm)	<i>T</i> (mm)
150	0.62	0.1	0.12
200	0.45	0.15	0.09

 Table 3.2: FSS parameters

wavelength slots are fabricated on both sides of a 0.4 mm quartz wafer to result in a second-order filter response with sharper shape factor as compared to the single resonator [53]. The slot length has a length of $L = (1/\sqrt{\epsilon_m}) \times \lambda_0/2$ where $\epsilon_m = (1 + \epsilon_r)/2$ and $\epsilon_r = 3.8$. The infinite 2-D array of unit cells was synthesized in Ansys-HFSS [31] through the use of perfect electric conductor (PEC) and perfect magnetic conductor (PMC) walls around a single unit cell. The resulting two FSS designs are summarized in Table 3.2.

The quasi-optical network analyzer is assembled as shown in Fig. 3.18. The FSS



Figure 3.18: FSS measurement setup.

was positioned halfway between the transmitter and receiver at a distance of 30 cm. The FSS appears in the far-field of both antennas resulting in an RF wave with an equi-phase front at the surface of the FSS ($R_{far-field} = 2D^2/\lambda = 28.8$ cm at 300 GHz for D = 12 mm). The measurements of the transmission coefficient (S₂₁) are done by measuring the (S₂₁) with and without the FSS, and the difference between the two cases is the FSS frequency response. The measured and simulated (S₂₁) of two FSS DUTs at 150 and 200 GHz are shown in Fig. 3.19.

The cross-slot is symmetrical and therefore has a substantial cross-polarized response (V-H). Since a polarizer is not used at the transmitter and receiver systems, any slight misalignment can result in a small cross-pol component which affects the rejection level. This is clearly seen in the cut-off region measurements of the FSS. The FSS response at 150 GHz and 200 GHz is well characterized by the 100 - 300 GHz quasioptical network analyzer.

The FSS can be modeled by two parallel LC tanks with a $\lambda/4$ between them. The measured 3-dB FSS fractional bandwidth can be used to calculate the filter loaded quality factor Q_L and therefore the lumped element values of the filter using

$$Q_L = \frac{1}{\Delta f_{3dB}} \tag{3.11}$$

$$L = \frac{\eta_0}{2\omega_0 Q_L} \tag{3.12}$$

$$C = \frac{1}{\omega_0^2 L} \tag{3.13}$$

The 150 GHz filter results in a loaded Q of 6.25, L = 32 pH, and C = 35.2 fF while the 200 GHz filter results in a loaded Q of 9.52, L = 15.76 pH, and C = 40 fF. A 410 μ m empty silicon wafer is also tested using the quasi-optical network analyzer to measure



Figure 3.19: Measured and simulated FSS insertion loss in (dB) at: (a) $f_0 = 150$ GHz with different orientations, (b) $f_0 = 200$ GHz.


Figure 3.20: Measured and simulated insertion loss of a 410 μ m empty silicon wafer.

the silicon insertion loss and shown in Fig. 3.20.

3.4 Conclusion

A 100 – 300 GHz quasi-optical network analyzer using compact Schottky-diode harmonic transmitter and receiver modules is demonstrated. The measured system EIRP and conversion gain agree well with simulations. The system is compact with a planar architecture and can be scaled to THz frequencies with the use of a smaller bow-tie antenna (300 GHz to 1200 GHz) and monolithically integrated diodes. The frequency response of two different FSS filters at 150 GHz and 200 GHz agree well with simulations. It is expected that better performance and higher dynamic range can be achieved using polarizers and a Gaussian-beam waveguide system between the transmitter and receiver units.

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Chapter 4

A Tunable Dual-Band Quadruple-Pole LTE Antenna for Carrier Aggregation Systems

4.1 Introduction

The exponential increase of traffic in mobile communication networks due to the surge in demand for accessing mobile data drove the research to the development and design of next generation mobile communication systems. Multiband antennas based on the planar inverted-F antenna (PIFA) have been greatly used in modern wireless devices such as smart mobile phones [54, 56–66, 90]. PIFAs are compact in size and can easily designed to resonate at multiple frequencies within 0.7 - 2.7 GHz range [54, 56–63, 90]. These antennas can also designed with multiple resonances resulting in wideband operation for different applications [64–66].

Fourth generation (4G) systems require a large bandwidth to achieve the targeted high data-rates. Carrier aggregation (CA) is established as a key enabler of LTE- Advanced to meet a peak data rates of up to 1 Gbps/s [67–69]. In the first implementation of the CA standard, two receive frequencies are used to improve the download data rate. One of the receive frequencies is always paired with the transmit frequency (as a standard W-CDMA pair), and the other frequency can be located anywhere. In the second implementation of the CA standard, three receive frequencies are used, one for transmit frequency and two for receive frequencies, or vise versa. In the third implementation of the CA standard, four receive frequencies are used to improve the download data rate, two for transmit frequency and two for receive frequencies, as shown in Fig. 4.1(a). In this case two resonances are located in the lower LTE band (0.7 - 1 GHz) and two at the higher LTE band (1.6 - 2.6 GHz).

Varactor diodes are variable capacitors that are commonly used as tuners in different RF applications. Even with their low Q (30-50) which results in relatively high loss designs, they are cost effective and easily controlled by a DC voltage. They still provide relatively good antenna efficiency (25-65%). The PIFA antennas can also be loaded with High-Q RF-MEMS variable capacitors (Q > 200) to result in tunable operation, thereby optimizing the antenna radiation efficiency for different bands.

This chapter presents a dual-feed, dual-band antenna with quadruple resonances covering the 0.7 - 1 GHz and 1.6 - 2.6 GHz bands. The antenna is loaded with four varactor diodes and results in independent tuning to meet the requirements of carrier aggregation systems. The antenna is designed to provide high isolation between the two different arms. The application areas are in carrier aggregation systems for 4G wireless systems.



Figure 4.1: (A) Frequency spectrum of quadruple-pole antenna for CA systems. (b) wideband response with a third tuner, (c) layout of the dual-feed quadruple-pole antenna.

4.2 Theoretical Analysis

4.2.1 Antenna Design

An antenna with two feeds loaded with four different series varactor diodes D_1 , D_2 at low-band and D_3 , D_4 at high-band is shown in Fig. 4.1(c). The feeds are superseded by a ground plane and shorting pins are used to isolate the feeds from each other. Each varactor diode is a variable capacitor and will force the antenna to resonate at a certain frequency (F_{C1} to F_{C4}) as shown in Fig. 4.1(a) creating four different resonances. Each resonance is independently tunable within a frequency range and to be used for transmit and receive. This antenna provides a wideband performance for higher data-rate to satisfy the high demand of speed in advanced communication systems.

There are important parameters that should to be considered to design tunable antennas. These parameters include: the antenna size, the tuner position, the tuner capacitance range, the spacing between tuners, and the quality factor of the tuner. Each parameter will be analyzed separately to see its effect on the antenna and then to optimize it to get better performance. The schematic of unloaded PIFA is shown in Fig. 4.2(a). In general for PIFA, the antenna length should be a quarter wavelength to resonate. The first resonance will occur when $Z_P = 0$ where the antenna feed should to be connected between the short end and the center of the antenna and its location is optimized to match the antenna to 50 Ω .

A PIFA loaded by a variable series capacitor C_S (varactor diode) is shown in Fig. 4.2(b) to tune its resonant frequency. The series capacitor will behave as a negative transmission line where its length is determined by its capacitance. This will reduce the antenna overall length and therefore force the antenna to resonate above its original resonant frequency. This topology has been already analyzed by [65, 66] showing that



Figure 4.2: A schematic of: (a) unloaded PIFA, (b) PIFA loaded with series capacitor, (c) PIFA loaded with two series capacitors w/o shunt capacitor.

$$Z_1 = \frac{Z_{o1}}{j\tan(\beta L_{new1})} \tag{4.1}$$

$$L_{new1} = \frac{1}{\beta} \tan^{-1} \left(\frac{Z_{o1} \omega C_S \tan(\beta L_1)}{\tan(\beta L_1) + Z_{o1} \omega C_S} \right)$$
(4.2)

where the new length can be controlled by Z_{o1} , the characteristic impedance of the line, the value and the position of the variable series capacitor C_S where β is the propagation constant.

A new concept to achieve wideband response is shown in Fig. 4.2(c), where the antenna is loaded by two different variable series capacitors C_1 and C_2 . These two capacitors can also behave as two different negative transmission lines forcing the antenna to resonate at two different resonant frequencies and their lengths depend on their capacitances. This circuit can be analyzed similarly with input impedance Z_x and two different new lengths L_{new1-2} where

$$Z_x = f(Z_{o1-3}, L_{new2-3}, C_1, C_2)$$
(4.3)

The antenna will have two different paths from the short to open ends forcing it to resonate at two different frequencies. These two resonances can be controlled by the



Figure 4.3: Effects of tuner (a) size, (b) position, on the center resonant frequency and tuning range using a capacitance of 1 to 9 pF.

antenna size (characteristic impedance and length), the value and the position of C_1 and C_2 . These three parameters can determine where the tuning center frequency f_{C1} and f_{C2} are and how much tuning range the antenna will have.

A study on the effects of the size and position of the tuner are shown in Fig. 4.3(a-b). The circuit in Fig. 4.2(b) is simulated using Agilent-ADS [32]. A set of capacitance range of 1 pF $\leq C_S \leq$ 9 pF is used to find f_{min} and f_{max} and to calculate the center frequency (f_c) and the tuning range (T) where

$$f_c = \sqrt{f_{max} f_{min}} \tag{4.4}$$

and

$$T = \frac{f_{max} - f_{min}}{f_{center}} \tag{4.5}$$

The characteristic impedance of the transmission lines is chosen to be 100 Ω .

The analysis shows that the tuning range will increase with the antenna size for the same capacitance range. The reason for that is that the increased antenna size will give the series capacitor the area of freedom to tune the antenna to higher frequency range as shown in Fig. 4.3(a) where the antenna size is in air. Therefore, a loaded antenna with series tuner has to be larger than its original size to achieve wideband tuning range result in a trade-off between the size and the tuning range. However, the overall operating frequency will shift down with increasing the tuning range.

The position of the tuner has also a strong effect on the tuning range. It is defined to be the ratio between the distance from the tuner and the open end L_1 to the total length L_T . The analysis shows that when the capacitor is approaching towards the short end, the length between the tuner and the open end L_1 will be larger giving the tuner more area to affect the antenna length and control its resonant frequency location and therefore enhance the tuning capability as shown in Fig. 4.3(b). The analysis also shows that at a certain position, the antenna will reach its tuning limit. However, the center frequency f_c will be shifted to higher frequency with increasing the tuning range. When the variable capacitor approaches the short end, it will approach the feed line too and this will complicate the input matching for the wide frequency range. In this design, both the antenna size and the tuners position are designed for good input matching and higher tuning range.

The quality factor of the tuner (Q) plays an important role in determining the performance of the antenna. In general, the tuner quality factor describes how much loss inside the device. A low Q can extremely affect the antenna efficiency and results

in poor radiation patterns. The Q of a series capacitor can be calculated by

$$Q = \frac{1}{2\pi f_C C_S R_s} \tag{4.6}$$

where f_C is the resonant frequency, C_S is the series capacitance, and R_s is the tuner's equivalent series resistance.

Ideally, to avoid the losses of the tuners, their Q should be infinity. Since the antenna resonant frequency range is specified to be $f_{min} \leq f \leq f_{max}$, the quality factor can be maximized by reducing the capacitance and the equivalent series resistance. As mentioned before, the capacitance can be controlled by the size of the antenna and the tuner position. One technique to minimize the values of capacitors used - to maximize their quality factor - is to choose the length of the antenna such that its maximum resonant frequency (f_{max}) occurs at C = 0 (ideal case) where the quality factor of the tuner is infinite. This will occur when

$$L_2 = L_T - L_1 = \frac{\lambda}{4} \quad at \quad f = f_{max} \tag{4.7}$$

and at $C_S = 0$, the tuner will be an open circuit and the antenna will resonate at f_{max} . With the varactor diodes which have a C_{min} , the previous method can be used where f_{max} will occur at $C_S = C_{min}$. The second factor that affects the Q is the equivalent series resistance R_s of the tuner. This resistor describes the energy loss from the tuner which will affect the antenna radiation pattern and efficiency too. Low capacitance and low equivalent series resistance are the keys to design a tunable antenna with different resonances at the same time with good radiation patterns and high radiation efficiency.

A study on the effect of the Q on (PIFA) radiation efficiency is shown in Fig. 4.4(a). The antenna is simulated using Ansys-HFSS [31] to resonate at f = 0.8 GHz and radiation efficiency is extracted with different Q (R_s changes). The simulation shows



Figure 4.4: (a) Tuners quality factor effect on antenna efficiency, (b) selection of capacitor values with different tuners spacing.

that with low Q (< 30), the radiation efficiency can be < 40%. The efficiency starts to increase as Q increases and it can reach 70% with high Q resonators such as RF-MEMS (Q > 200). After Q of 250, the antenna efficiency becomes limited by the highly loaded components and conductor/dielectric losses which are also factors that degrade the antenna radiation efficiency.

The spacing distance (*S*) between the two variables capacitors (C_1) and (C_2) in Fig. 4.2(c) are analyzed to see its effects on capacitance range. In this design, the antenna has to resonate at close frequencies to double the bandwidth as shown in Fig. 4.1(b) and the frequency is chosen to be (f = 0.84 GHz). The analysis shows that a small spacing will require high capacitances to force the antenna to resonate. The spacing between the tuners is chosen to be 9 mm to reduce the capacitances for higher Q and to give an area of freedom for C_1 and C_2 to tune to low and high frequencies respectively (Fig.4.4(b)).

4.2.2 Wideband Response with a Third Tuner

A novel technique to achieve wider bandwidth is to add another tuning property. Previously, the two series capacitors are tunable within a certain frequency range and they can only merge together (for $2 \times BW$) at one frequency which is chosen to be at the med band. A new way to tune the merged resonant frequencies within the whole LTE range is introduced .In Fig. 4.2(c), a shunt capacitor (C_M) is connected to the shorting pins of the antenna in order to tune the two original resonances together. After adding the branch of CM, the input impedance (Z_Y) of the branch is

$$Z_{Y} = Z_{o4} \left(\frac{\frac{1}{j\omega C_{M}} + jZ_{o4} \tan(\beta L_{4})}{Z_{o4} + j\frac{1}{j\omega C_{M}} \tan(\beta L_{4})} \right)$$
(4.8)

 Z_Y can be also rewritten as open-terminated transmission line with characteristic impedance of Z_{o4} , and propagation constant β , and length L_{new4} where

$$Z_Y = \frac{Z_{o4}}{j\tan(\beta L_{new4})} \tag{4.9}$$

$$L_{new4} = \frac{1}{\beta} \tan^{-1} \left(\frac{Z_{o4} \omega C_M + \tan(\beta L_4)}{1 - Z_{o4} \omega C_M \tan(\beta L_4)} \right)$$
(4.10)

Tuning C_M will also tune the combined resonant frequency (f_{CW}) to low LTE frequency region range from 1 to 0.7 GHz as shown in Fig. 4.5(a). However, the shunt capacitance will behave as a positive transmission line and therefore it will increase the size of the antenna and therefore tuning it to lower frequencies. By choosing the antenna size and the position of the tuners carefully, we can push the two resonances to be very close to each other and combine them to achieve wideband response centered at f_{CW} (the center frequency of the combined two resonances). Without (C_M), f_{CW} is chosen to be closer to the higher edge of the frequency range (around 0.95 GHz). After adding



Figure 4.5: (a) Simulated S_{11} with different C_M values at low-band, (b) effect of C_M on center resonant frequency.

and tuning C_M , The f_{CW} will be tuned up and down between 0.725 – 0.95 GHz. The location of C_M should be chosen carefully to cover the overall LTE frequency range.

Another advantage of C_M is that it can be used to shift the harmonics produced by the low frequency antenna (0.7 – 1 GHz) and prevent any interruption to any close antenna that resonates at (f = 1.6 - 2.6 GHz). This will allow building multiple antennas close to each other and with high isolation between them. In this chapter, two antennas will be built next to each other to cover the whole LTE frequency band (0.7 – 2.6 GHz). The required shunt capacitance (C_M) for both antennas is tested and shown in Fig. 4.5(b). To tune f_{CW} from 0.95 GHz to 0.72 GHz, C_M should be tuned from 0 to 1.6 pF for the low-band antenna and it is slightly higher (up to 2 pF) to tune f_{CW} from 2.3 GHz to 1.7 GHz in high-band antenna.

4.3 Dual-feed Dual-Band Quadruple-Pole Antenna

4.3.1 Antenna Design

An antenna with two feeds and four series variable capacitors as tuners is shown in Fig. 4.1(a). The properties of such an antenna can be used for carrier aggregation and MIMO system applications. The antenna is generally a combination of two antennas next to each other. The first antenna is designed to be tuned with two resonant frequencies within the low LTE band (0.7 – 1 GHz) while the second antenna is designed to be tuned with two resonant frequencies within the high LTE band (1.6 – 2.6 GHz) making it a quadruple-pole antenna. Shorting pins to ground are used between the two parts to provide the ground requirement of PIFA and to enhance the isolation between them. The antenna size is 36 - 18 mm and it is printed on $66 \times 100 \times 3.154$ mm circuit board using 2-ply 60 mils Rogers RO4003C ($\varepsilon_r = 3.55$, $tan\delta = 0.0027$ at 2.5 GHz) stacked together using 1-ply 4 mils Rogers RO4450B ($\varepsilon_r = 3.54$, $tan\delta = 0.003$ at 10 GHz).

Each antenna is loaded with two different variable tuners (varactor diodes) to resonate at four different frequencies at the same time providing a quadruple-pole response. Four different DC voltages (V_1 , V_2 , V_3 , and V_4) are applied to the diodes separately to tune the their capacitances respectively. The first resonant frequency is tunable using a Skyworks diode SMV-1235-SC-79, with $C_t = 2.38 - 18.22$ pF, $R_s = 0.6 \Omega$, and $V_{Bias} = 0.15$ V. The second resonant frequency is tunable using a Skyworks diode SMV-1234-SC-79, with $C_t = 1.32 - 9.63$ pF, $R_s = 0.8$, and $V_{Bias} = 0.15$ V. The third and four resonant frequencies are tunable using Skyworks diodes SMV-1405-SC-79, with $C_t = 0.63 - 2.67$ pF, $R_s = 0.8$, and $V_{Bias} = 0.30$ V. A high capacitance (C_b) > 50 nF is connected between the two diodes in each antenna to block the DC voltage and isolate the diodes from each other. The antenna is fed using coaxial cables through SMA connectors attached to the antenna ground. The location of the feeds is optimized for wideband



Figure 4.6: Dual-feed quadruple-pole antenna photograph.



Figure 4.7: Lumped circuit model for the dual-feed dual-band quadruple-pole antenna.

input matching. The ground plane is slotted to help in both matching and radiation. Photograph and circuit model of the antenna are shown in Figs. 4.6 and 4.7.

4.3.2 Input Matching, Isolation, and Tuning Capabilities

The measured antenna input S-parameters (S₁₁ and S₂₂) are shown in Fig. 4.8(ab). The low-band antenna has two independently tunable resonant frequencies. The first resonance is tunable from 0.7 - 0.85 GHz using a capacitance of 2.38 - 7 pF and DC voltage 2.6 - 15 V while the second resonance is tunable from 0.86 - 1 GHz using a capacitance of 2.4 - 4.68 pF and DC voltage 2 - 4.9 V. The two resonances are combining together at center frequency of 0.86 GHz providing a wideband matching response and enhancing the design performance. This wideband response can be tunable too using a shunt capacitor connected to short pins. The high-band antenna has two independently tunable resonant frequencies. The first resonance is tunable from 1.6 - 2 GHz using a capacitance of 1.75 - 2.6 pF and DC voltage 1.5 - 4.25 V while the second resonance is tunable from 2 - 2.6 GHz using a capacitance of 0.77 - 2.2 pF and DC voltage 0.4 - 20 V. The two resonances are combining together at center frequency = 2 GHz providing a wideband matching response.

The measured isolation coefficients (S_{12} , S_{21}) between the antennas are shown in Fig. 4.9(a-b). The effect of antenna 2 (high-band) on antenna 1 (low-band) is less than -22 dB all over the frequency range providing a high isolation in the low-band. The effect of antenna 1 on antenna 2 is less than -13 dB all over the frequency range providing a good isolation at low-band. The isolation in the high-band is degraded due to contributions from the second and third harmonics produced by the low-band antenna. One way to suppress the harmonics is to add a notch filter to remove all the high-band harmonics radiated from the low-band antenna. The linearity of these antennas was not tested but in similar designs using Skyworks diodes, the expected third order Inputintercept point (*IIP*₃) can be around 16 – 24 dBm. A better way to improve the linearity is to use high DC bias voltages to tune the varactor diodes where the RMS voltage from high input power has small effects.

The peak voltages across the all the tuners are simulated in Fig. 4.10 to see the voltage swing across the varactor diodes and how can affect the antenna linearity. The simulated voltages are achieved by applying a 20 dBm input power at each frequency. The peak voltage is around 5 - 15 V and 10 - 17 V across the first and second resonances respectively. In the high-band antenna, the peak voltage across the third resonance is between 10 - 15 V and around 3 - 12 V across the fourth resonance. The AC analysis is done using Ansys-HFSS and Agilent-ADS software tools. The voltage is high with $P_{in} = 20$ dBm and it will cause very non-linear components at these power levels and ending up with frequency shifting. Fortunately, RF-MEMS can handle this levels of power changes ($V_{max} = 40$ V) for Cavendish Kinetics variable capacitance de-



Figure 4.8: Measured and simulated S-parameters (in dB): (a) S₁₁, (b) S₂₂.



Figure 4.9: Measured isolation S-parameters (in dB): (a) S₁₂, (b) S₂₁.



Figure 4.10: Voltage across tuners when $P_{in} = 20$ dBm.

vices [12] and with a Q > 200, the performance of the antenna will be highly improved.

4.3.3 Radiation Patterns and Antenna Efficiency

The radiation patterns and the efficiency of low-band antenna are presented in Figs. 4.12-4.13. The measurements are done inside Satimo Stargate-32 chamber in Fig. 4.11 [71]. The measured radiation patterns are shown at 0.79 and 0.92 GHz in XY and YZ planes. The total patterns have an isotropic radiation as expected from PIFA. Small variations appeared in the patterns due to input cable radiation. The simulated and measured efficiency agree well and they are around 25-50% using Skyworks diodes. The simulated efficiency using RF-MEMS (Q > 200) is higher than 50% which proves that the radiation efficiency is limited by the quality factor of the tuners. Other reasons of efficiency degradation include the highly loaded antenna, conductor, and dielectric losses.

The radiation patterns and efficiency of the high-band antenna are presented in Figs. 4.14-4.15. The measured radiation patterns are shown at 1.8 and 2.3 GHz in XY and YZ planes. The total patterns have an isotropic radiation as expected from PIFA. The radiation due to the input cable seems to be smaller in this antenna. The simulated and measured efficiency agree well and they are around 30-62% using Skyworks diodes. The simulated efficiency using RF-MEMS (Q > 200) is higher than 50% which proves



Figure 4.11: A reference photograph of Satimo Stargate-32 chamber for pattern measurements. [71]



Figure 4.12: Measured low-band antenna radiation patterns.



Figure 4.13: Measured and simulated low-band antenna radiation efficiency.



Figure 4.14: Measured high-band antenna radiation patterns.



Figure 4.15: Measured and simulated high-band antenna radiation efficiency.

that the radiation efficiency is limited by the quality factor of the tuners. Other reasons of efficiency degradation include the highly loaded antenna, conductor, and dielectric losses.

4.4 Wideband Response with the Third Tuner

The antennas with an extra shunt variable capacitor (C_M) are shown in Fig. 4.1(c). The shunt capacitor as mentioned will add more length to the antenna and tuning the f_{CW} to lower frequency range. The original two series tuners in each antenna are relocated to combined at the higher side of the band at 0.95 GHz and 2.3 GHz at low-and high- bands antennas respectively. With tuning C_M up to 2 pF, the combined wide-

band resonance will be tunable within the whole LTE band as shown in Fig. 4.16(a-b). Before adding C_M , the measured bandwidth is 20 – 25 MHz at the low-band antenna where the bandwidth is 60 – 70 MHz at the high-band antenna. With C_M , the measured bandwidth increases to 40 – 50 MHz at the low-band antenna where the bandwidth increases to 120 – 150 MHz at the high-band antenna. Despite the fact that adding another varactor diode will affect the antenna radiation efficiency, since it has losses due to series resistance, the wider bandwidth obtained proved worthwhile. Instead of using the varactor diode C_M and since we only need several capacitances to cover the whole LTE band (only 7 – 10 different capacitances for each shunt tuner), switches like SP7T or SP11T connected to several fixed capacitors can be used. This will enhance the design with high Q components and therefore, it will improve the antenna radiation efficiency (>50%).

The idea of combining the two resonant frequencies to increase the bandwidth is simulated in head and hand environment to check its reliability in real applications. This analysis is done using Ansys-HFSS. The antenna is placed adjacent to a model of a human head and hand to see the antenna performance in practical environment. Due to the dielectric differences between the human body and air, a degradation in both antenna input return loss and its radiation efficiency is expected. The head and hand are placed close to the antenna with a gap < 1 cm as shown in Fig. 4.17(a). The antenna return loss is still less than -7 dB. With minor design modification, the matching is optimized to be less than -10 dB as shown in Fig. 4.17(b). The modification will not affect the antenna performance if we remove the head and hand. The radiation efficiency with head and hand.



Figure 4.16: Measured S-parameters (in dB) with different C_M : (a) S_{11} , (b) S_{22} .

4.5 Conclusion

In this chapter, a study of the effects of loading an antenna with set of capacitance is presented. An LTE antenna with dual feeds is presented. The antenna has the capability of tuning at multiband frequencies, thus supporting carrier aggregation and providing a wideband performance to satisfy the requirements of modern LTE standards. The design is a dual-feed quadruple-pole antenna with two tunable resonances at low-band (0.7 - 1 GHz) and two tunable resonances at high-band (1.6 - 2.6 GHz). The antenna radiation patterns are isotropic as we expected from PIFAs. The radiation efficiency is between 25-60% using varactor diodes and higher than 50% when higher quality factor tuners are used such as RF-MEMS (Q > 250 at 0.7 GHz). A new method is introduced to increase the bandwidth by adding a shunt capacitor. This new antenna is simulated inside a head and hand environment and results in a good performance making the design reliable for real applications.

4.6 Acknowledgment

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Figure 4.17: (a) Head and hand setup [72], (b) simulated S_{11} .

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Chapter 5

RF MEMS Tunable 4-Pole Bandpass Filters With Bandwidth Control and Improved Stopband Rejection

5.1 Introduction

There have been substantial efforts in the past few years to develop new techniques to build high performance tunable RF filters for modern multi-standard communication systems [73]. Microstrip combline tunable filters are commonly used due to their compact size, and tunability which can be obtained using Schottky diode varactors [74–85]. However, these diodes have low Q (30-50) which results in relatively high loss designs and most of them cannot meet the linearity or power handling requirements of modern communication systems. Recently, RF MEMS (radio frequency micro-electro-mechanical-systems) devices are attracting a lot of attention due to their high-Q, high linearity, large power handling, small size, and low power consumption, all making them ideal elements to build high performance tunable filters in the 0.5-6



Figure 5.1: (a) Circuit model of the proposed filter structure, and (b) equivalent circuit of the coupled lines.

GHz range. This has led to several demonstrations of tunable filters with RF MEMS devices [86–91]. However, all prior work used in-house MEMS devices, which are not packaged and suffer from reliability concerns.

RF-MEMS technology is experiencing a huge development effort to enhance its capabilities and commercial vendors are now producing MEMS packaged devices with high reliability. This include Cavendish Kinetics and wiSpry for capacitive switches [70,92], and other manufacturers are working on high-reliability ohmic contact switches [93,94]. These devices are small ($\approx 2 \text{ mm}^2$ bumped die), hermetically-packaged, and ESD protected with excellent long-term reliability under medium to high RF power conditions (up to 2 W). Cavendish Kinetics switched capacitors, for example, can handle RF power of 33 dBm with cycling > 10B cycles and switching time < 50 μ s. They are also easy to control with SPI or RFFE digital controls. Several designs were recently demonstrated using these devices, in particular, RF filters and phase shifters [95, 96]. Their results show a substantial improvement in terms of insertion loss, linearity, and power handling.

Several attempts to improve the stopband rejection of tunable filters are presented [83–85]. Three and four-pole tunable filters with extra transmission zeroes were demonstrated in [83,85]. These zeroes are located close to the passband to improve the filter stopband rejection and enhance the filter selectivity. Others techniques to tune the filter's passband bandwidth are also demonstrated [76, 84–86].

This paper, a continued work of [85, 97], presents tunable 4-pole bandpass filters using commercial RF-MEMS capacitors and varactor diodes 0.7 - 2.4 GHz applications. The work presents low-loss and highly linear tunable filters with bandwidth control and improved stopband rejection. The work also discusses the different requirements of SPI and RFFE digital controls.

5.2 Filter Design

The proposed tunable 4-pole band-pass filter is shown in Fig. 5.1(a). This filter basically consists of three $\lambda_d/4$ combline resonators loaded by RF-MEMS Cavendish Kinetics varactors, DVC_1 - DVC_3 . The opposite ends with via holes are modeled by small-value inductors instead of ideal short circuits. Via-hole inductance determines filter bandwidth if band rejection is required to be the same and under same rejection, filter with larger via-hole inductance (narrower diameter or thicker substrate) has wider bandwidth. The input and output coupling is achieved by a tapped-line connection, and the varactors C_{IN} are used to obtain an impedance match over a wide frequency range. The coupling values between the adjacent resonators include the intrinsic coupling of the coupled lines, which are equivalent to the circuit shown in Fig. 5.1(b), and the capacitive coupling introduced by varactors C_{12} and C_{23} . For the non-adjacent resonators, there is an additional coupling path consisting of transmission lines and varactor C₁₃ for the purpose of generating extra transmission zeroes. The simulated (S_{21}) with and without the non-adjacent resonator is shown in Fig. 5.2 and done using Sonnet and Agilent-ADS [32, 98]. Moreover, this coupling path can be seen as a resonator (a back-to-back combline resonator) which is intentionally designed to have a resonant frequency at the



Figure 5.2: Filter response with and without the non-adjacent coupling path.

passband. As a result, A four-pole filter response is obtained with four transmission zeroes to improve the stopband rejection.

The resonant frequency of the combline-line resonator is determined by the characteristic impedance and the length of the transmission line in addition to the loaded capacitor. The length of the resonators are the same in the design while the impedances Z_1 , Z_3 and the states of DVC₁ and DVC₃ are the same for symmetrical performance. If the electrical length is short and less than 30° at ω_0 , the resonant frequency can be approximately calculated by [99]

$$\omega_{0i} \approx \sqrt{\frac{\omega_0}{Z_i \theta_i C_i}}, \quad i = 1, 2, 3 \tag{5.1}$$

$$\omega_{013} \approx \sqrt{\frac{\omega_0}{2Z_{13}\theta_{13}C_{13}}}$$
 (5.2)

The coupling coefficients between the resonators which affect the bandwidth of the filter can be calculated using

$$K = \frac{\omega_{0e}^2 - \omega_{0o}^2}{\omega_{0e}^2 + \omega_{0o}^2}$$
(5.3)

where ω_{0e} is the even-mode resonant frequency as the expression in (5.1) and



Figure 5.3: Layout of the 0.7 - 1 GHz tunable 4-pole band-pass filter.

 ω_{0o} is the odd-mode resonant frequency given by

$$\omega_{0o} \approx \sqrt{\frac{\omega_0}{(Z_1//\frac{Z_{12}}{2})\theta_1(C_1 + 2C_{12})}}$$
(5.4)

The coupling coefficient of the resonator in the auxiliary path is

$$K_{aux} \approx \frac{\omega_0^2 L_{via}}{\sqrt{\frac{Z_1 Z_{13} \theta_1 \theta_{13}}{2C_1 C_{13}}}}$$
 (5.5)

The external Q is evaluated by

$$Q_e \approx \frac{R_0 \omega_0}{\omega_{01} Z_1 \theta_T} \approx \frac{R_0}{\theta_T} \sqrt{\frac{\omega_0 C_1 \theta_1}{Z_1}}$$
(5.6)

The equations of the transmission zeroes which can be obtained by even- and odd-mode analysis are rather complicated and thus are not addressed in this paper. Instead, the investigation of zero positions by changing the capacitors C_{12} (and C_{23}) is done in [85]. It is seen that by increasing the value of C_{12} , the zeroes at the higher/lower frequencies will be pushed toward/away from the passband. Also, the passband bandwidth varies with C_{12} . The proposed filter can be used in a variety of wireless communication systems due to its capability of bandwidth and transmission zero control.



Figure 5.4: CK device with SPI control: (a) block diagram, (b) capacitance and quality factor at 0.85 GHz.

5.3 0.7 - 1 GHz Tunable 4-Pole Bandpass Filter with SPI Digital Control

A tunable 4-pole band-pass filter is designed to cover a frequency range of 0.7 to 1 GHz. The filter layout is shown in Fig. 5.3. The filter is printed on 50 mils printedcircuit board using Rogers RT/duroid 6010LM ($\varepsilon_r = 10.7$, $tan\delta = 0.0023$ at 10 GHz) and soldermask is used to cover the transmission-lines in order to reduce the loss introduced by electroless nickel immersion gold (EPIG) metal finish [95]. The length of the combline resonators are 9.6 mm (23° at $f_{0,max}$) and allow the use of equations (5.1 and 5.2) to approximate the operating frequency of the filter while the characteristic impedance of the resonators is $\approx 50 \ \Omega$. Three hole-vias with radius of 0.2 mm are located between each two resonators to provide a ground connection and allow to use $\lambda/4$ combline resonators. Large vias are designed to have smaller inductance and therefore to have narrower bandwidth with the same band rejection.

The main three resonators are tunable using Cavendish Kinetics CK505S ca-

pacitor. Each resonator is loaded in parallel with a 2 pF high-Q ATC capacitor to add more capacitance. The 5-bit Cavendish Kinetics RF-MEMS Digital Variable Capacitor (DVC) is built on a 0.18 μ m CMOS platform with chip scale hermetic packaging and flip chip bumps, as shown in Fig. 5.4(a) [70]. The overall size of the device is 1.4 imes 1.3 mm², which include the RF-MEMS capacitor banks at the bottom and the control circuits and charge pump. The device requires only 1.8 V to operate and is easily integrated into existing communication systems. A capacitance value of 1.1-5.1 pF is achieved over 32 states with a capacitance step of 125 fF. The device is controlled by serial peripheral interface (SPI) where three digital lines (CLK, SDA, and CSN) are required. A mechanical switch is placed on the path of CSN signal to control the state of each device individually. The switch can be turned on to change the state of one device while keeping the other two at the same state. The measured device capacitance and Q at 0.85 GHz are shown in Fig. 5.4(b). The measured device Q is 434-90 from state 0 to state 31, respectively, at 0.85 GHz, while the capacitance has an extra 0.1 pF from the model. The measured IIP3 of these devices is 65-67 dBm in a 50 Ω set-up over the entire capacitance range.

The coupling between the resonators (C₁₂, C₂₃) is tunable using a Skyworks diode SMV-1232-079LF, with $C_{12,23} = 0.72 - 4.15$ pF, $R_s = 1.5 \Omega$, L = 0.7 nH, and $V_{Bias} = 0.15$ V. The coupling devices are connected in series with 4.3 and 3.9 pF high-Q ATC capacitors, respectively to block the DC bias of the diodes and to control the capacitance range. The auxiliary resonator (C₁₃) is tunable using a Skyworks diode SMV-1413-079LF, with $C_{13} = 1.77 - 9.24$ pF, $R_s = 0.35 \Omega$, L = 0.7 nH, and $V_{Bias} =$ 0-30 V. A 100 pF high-Q ATC capacitor is connected in series with the varactor diode to block the DC bias of the diode. Two varactor diodes (C_{IN}) are connected to the input and output of the filter in series with 27 pF high-Q ATC capacitor to improve the matching. These diodes are Skyworks diodes SMV-1235-079LF, with $C_{IN} = 2.38 - 18.22$ pF, R_s



Figure 5.5: 0.7 - 1 GHz filter S-parameters: (a) S_{21} , (b) S_{11} .

= 0.6 Ω , L = 0.7 nH, and V_{Bias} = 0-15 V. Resistors of 30 k Ω are used to prevent any RF leakage to the DC sources.

5.3.1 S-Parameters

Fig. 5.5 presents the measured filter S-parameters (S_{21} , S_{11}) of the 0.7 – 1 GHz proposed filter. The measured filter tuning range is 0.7 to 1 GHz and successfully achieved using the 32 device steps (Fig. 5.6(a). The insertion loss is 2.9-2.6 dB with a 1-dB fractional bandwidth (FBW) of 6-8% (45-75 MHz). The measured S_{11} is < -18 dB for all states. Two transmission zeroes are present at both sides of the passband and provide a wideband stopband rejection which is > 33 dB. Measurements agree very well with simulations (not shown for brevity). The measured loaded-Q of a single combline-resonator with CK505S is 80-112 at 0.85 GHz.



Figure 5.6: 0.7 - 1 GHz filter: (a) center frequency vs. DVC state, (b) measured 1-dB bandwidth and loss vs. center frequency

5.3.2 Bandwidth Control and Shape Factor

The passband bandwidth of the filter can be tuned by controlling the coupling capacitor (C_{12}) as shown in Fig. 5.7(a). Tunable 1-dB FBW of 4.2-7.6% can be achieved with C_{12} DC voltage bias at 2.1-6.4 V. The selectivity of the bandpass filter is calculated from the S_{21} in terms of the shape factor (SF). The shape factor expresses the filter selectivity in another way and it can be calculated by determining the frequencies of the 6 and 30 dB points where

$$SF = \frac{F_{H,30dB} - F_{L,30dB}}{F_{H,6dB} - F_{L,6dB}}$$
(5.7)

The lower the shape factor, the steeper the skirts of the filter's response curve. Ideally, a shape factor of unity would indicate that the response curve of a filter has a rectangular shape. A shape factor of 1.5-1.7 is obtained for this filter which is enough for applications that employ narrowband wireless communication standards.



Figure 5.7: 0.7 - 1 GHz filter: (a) bandwidth control, (b) shape factor.

5.4 1.7 - 2.4 GHz Tunable 4-Pole Bandpass Filter with RFFE Digital Control

A tunable 4-pole band-pass filter is also designed to cover a frequency range of 1.7 to 2.4 GHz. The filter layout is shown in Fig. 5.8. The filter is printed on 50 mils printed-circuit board using Rogers RT/duroid 6010LM ($\varepsilon_r = 10.7$, $tan\delta = 0.0023$ at 10 GHz) and soldermask layer is used to cover the transmission-lines in order to reduce the loss introduced by electroless nickel immersion gold (EPIG) metal finish. The length of the combline resonators are 3.8 mm (21.8° at $f_{0,max}$) and the characteristic impedance of the resonators is $\approx 50 \ \Omega$. Three hole-vias with radius of 0.2 mm are located between each two resonators to provide a ground connection and allow to use $\lambda/4$ combline resonators (compact) and to have smaller inductance for narrower bandwidth and same band rejection.

The main three resonators are tunable using Cavendish Kinetics CK402R capac-



Figure 5.8: Layout of the 1.7 - 2.4 GHz tunable 4-pole band-pass filter.



Figure 5.9: CK device with RFFE control: (a) block diagram, (b) capacitance and quality factor.

itor. The 5-bit Cavendish Kinetics RF-MEMS Digital Variable Capacitor (DVC) is built on a 0.18 μ m CMOS platform with chip scale hermetic packaging and flip chip bumps, as shown in Fig. 5.9(a) [70]. The overall size of the device is 1.4 × 1.3 mm², which include the RF-MEMS capacitor banks at the bottom and the control circuits and charge pump. The device requires only 1.8 V to operate and is easily integrated into existing communication systems. A capacitance value of 0.62-2 pF is achieved over 32 states with a capacitance step of 43 fF.

Alternatively, this device is controlled by radio frequency front-end (RFFE) in-

terface. RFFE is a compact two-wire interface defining one line for system clock (CLK) and one line for the system data (SDA) where the SDA line is a bi-directional data signal that can control the slave state (CK device) and therefore, no slave select line (CSN) is needed. RFFE is a single-master system that avoids timing uncertainties inherent with bus arbitration and allow for bus operation up to 26 MHz [100]. RFFE employs a point-to-multipoint structure to enable control of multiple RF front-end devices. It also can support up to 15 slave devices on a single RFFE bus. 10 k Ω resistors are connected in the path of the clock and data lines and close to the device in order to limit any spurious coupling. A 100 pF bypass capacitor is connected to power supply pin to short any RF signals to ground and to avoid any RF noise that may be present on a DC signal. A mechanical switch is placed on the path of SDA signal to control the state of each device individually. The measured device capacitance and Q at 2 GHz are shown in Fig. 5.9(b). The measured device Q is 343-98 at 2 GHz from state 0 to state 31 while the capacitance has an extra 0.05-0.1 pF from the model. The measured IIP3 of these devices is 65-67 dBm in a 50 set-up over the entire capacitance range.

The coupling between the resonators (C₁₂, C₂₃) is tunable using a Skyworks diode SMV-2020-079LF, with $C_{12,23} = 0.35 - 3.2$ pF, $R_s = 2.5 \Omega$, L = 0.7 nH, and V_{Bias} = 0-20 V. The coupling devices are connected in series with 2 pF high-Q ATC capacitors to block the DC bias of the diodes and to control the capacitance range. The auxiliary resonator (C₁₃) is tunable using a Skyworks diode SMV-1405-040LF, with $C_{13} = 0.56 -$ 2.81 pF, $R_s = 0.75 \Omega$, L = 0.45 nH, and $V_{Bias} = 0.30$ V. A 2.7 pF high-Q ATC capacitor is connected in series with the varactor diode to block the DC bias of the diode and to control the capacitance range. Two varactor diodes C_{IN} are connected to the input and output of the filter in series with 27 pF high-Q ATC capacitor to improve the matching. These diodes are Skyworks diodes SMV-1235-079LF, with $C_{IN} = 2.38 - 18.22$ pF, R_s = 0.6 Ω , L = 0.7 nH, and $V_{Bias} = 0.15$ V. Resistors of 30 k Ω are used to prevent any RF



Figure 5.10: 1.7 - 2.4 GHz filter S-parameters: (a) S_{21} , (b) S_{11} .

leakage to the DC sources.

5.4.1 S-Parameters

Fig. 5.10 presents the measured filter S-parameters (S_{21} , S_{11}) of the 1.7 – 2.4 GHz proposed filter. The measured filter tuning range is 1.73 to 2.4 GHz and successfully achieved using the 32 device steps (Fig. 5.11(a). The insertion loss is 3.55-2.4 dB with a 1-dB fractional bandwidth (FBW) of 4.4-5.1% (85-105 MHz). The measured S_{11} is < -20 dB for all states. Two transmission zeroes are present at both sides of the passband and provide a wideband stopband rejection which is > 33 dB. Measurements agree very well with simulations (not shown for brevity). The measured loaded-Q of a single combline-resonator with CK402R is 86-113 at 2 GHz.


Figure 5.11: 1.7 - 2.4 GHz filter: (a) center frequency vs. DVC state, (b) measured 1-dB bandwidth and loss vs. center frequency

5.4.2 Bandwidth Control and Shape Factor

The passband bandwidth of the filter can be tuned by controlling the coupling capacitor (C_{12}) as shown in Fig. 5.12(a). Tunable 1-dB FBW of 4.1-7.4% can be achieved with C_{12} DC voltage bias at 3.9-9.4 V. The selectivity of the bandpass filter is calculated from the S_{21} in terms of the shape factor (SF). A shape factor of 1.59-1.65 is obtained for this filter which is enough for narrowband wireless communication applications.

5.5 Linearity and Complex-Signal Measurements

The measured gain of the 0.7 - 1 GHz filter versus the input power is shown in Fig. 5.13, at low, med, and high capacitive loading. The filter starts to saturate after 10 dBm for the high capacitive loading and results in 1-dB compression point (P1dB) of 16 dBm at 0.72 GHz (state 31) while it is approximately equal to 20 dBm for the lower loading at 1 GHz (state 0). The reason of the low P1dB is due to the high RF voltage swing across the varactor diodes resulting in high non-linear components. One special case with only RF-MEMS varactors and fixed capacitors at 1 GHz is demonstrated. The



Figure 5.12: 1.7 - 2.4 GHz filter: (a) bandwidth control, (b) shape factor.

1-dB compression point cannot be seen up to 26 dBm. The small gain variation of < 0.1 dB is due to the accuracy of the power meter.

The simulated root mean square (rms) voltages across the DVC terminals versus different state setting (or filter frequencies) for the 0.7 - 1 GHz filter are shown in Fig. 5.14(a) with an input power of 26 dBm. When the input power is 26 dBm, which is approximately 4 Vrms at the input of the filter, the rms voltage across the DVC is 6.3-11.6 V, which is at least 50% higher than the input voltage. The simulated rms voltages



Figure 5.13: Measured gain of 0.7 - 1 GHz filter vs. input power.



Figure 5.14: Simulated rms voltage of the 0.7 - 1 GHz filter across: (a) RF-MEMS devices with $P_{in} = 26$ dBm, (b) varactor diodes with $P_{in} = 10$ dBm.

across the varactor diode terminals versus different state setting (or filter frequencies) for the 0.7 - 1 GHz filter are shown in Fig. 5.14(b) with an input power of 10 dBm. The rms voltage across the diodes is 0.2-1 V. An rms voltage of 1 V at input power of 10 dBm can really affect the diode state and change the filter response, resulting in a lower IIP3.

The IIP3 of the filter is measured using the setup in Fig. 5.15. Two signals with a separation of 10 MHz are applied to the filter input where the fundamental and the third-order intermodulation products are measured (Fig. 5.16). An IIP3 of 25.2-28 dBm is obtained at low and high capacitive loading. These two states describe the best and worst performance cases. An IIP3 of 52 dBm can be obtained when only RF-MEMS varactors and fixed capacitors are used at 1 GHz. The measured filter IIP3 is lower than the device specification, which is 65-67 dBm in a 50 Ω setup.

A better way to show the effect of the filter non-linearity is to measure the spectral regrowth for a complex signal such as a 5-MHz wideband CDMA waveform (Fig. 5.17). It is seen that the ACPR is > 58.3 dB for an input power of 26 dBm at 1 GHz (state 0) with RF-MEMS and fixed capacitors, and the ACPR is 48-51 dB for the filter



Figure 5.15: Measurement setup for IIP3 for the tunable 4-pole band-pass filter.



Figure 5.16: (a) Measured IIP3 at 1 GHz with RF-MEMS and fixed capacitors, (b) measured IIP3 and P1dB for the 0.7 - 1 GHz filter with varactor diodes.



Figure 5.17: (a) Measured spectral regrowth at 26 dBm for a 5-MHz WCDMA signal at 1 GHz with RF-MEMS and fixed capacitors, (b) measured ACPR versus frequency for the 0.7 - 1 GHz filter with varactor diodes with $P_{in} = 10$ dBm.



Figure 5.18: Measured gain of the 1.7 - 2.4 GHz Filter vs. input power.

with varactor diodes for an input power of 10 dBm. In general, for W-CDMA signals, GaAs or CMOS cell phone power amplifiers have an ACPR of 36-40 dB, and therefore, the tunable filter will not contribute any additional distortion.

The measured gain of the 1.7 - 2.4 GHz filter versus the input power is shown in Fig. 5.18, at low, med, and high capacitive loading. The filter starts to saturate after 10 dBm for the high capacitive loading and results in 1-dB compression point (P1dB) of 13.5 dBm at 1.75 GHz (state 31) while it is approximately equal to 22 dBm at the lower loading at 1 GHz (state 0). The special case with only RF-MEMS varactors and



Figure 5.19: Simulated rms voltage of the 1.7 - 2.4 GHz Filter across: (a) RF-MEMS devices with $P_{in} = 26$ dBm, (b) varactor diodes with $P_{in} = 10$ dBm.

fixed capacitors at 2 GHz is demonstrated. The 1-dB compression point cannot be seen up to 26 dBm. The small gain variation of < 0.1 dB is due to the accuracy of the power meter.

The simulated rms voltages across the DVC terminals versus different state setting (or filter frequencies) for the 1.7 - 2.4 GHz filter are shown in Fig. 5.19(a) with an input power of 26 dBm. The rms voltage across the DVC is 6.8-18.5 V, which is at least 50% higher than the input voltage. The simulated rms voltages across the varactor diode terminals versus different state setting (or filter frequencies) for the 1.7 - 2.7 GHz filter are shown in Fig. 5.19(b) with an input power of 10 dBm. The rms voltage across the diodes is 0.1-1 V. An rms voltage of 1 V at input power of 10 dBm can really affect the diode state and change the filter response, resulting in a lower IIP3.

The measured IIP3 is 25.2-28 dBm is obtained at low and high capacitive loading. These two states describe the best and worst cases (Fig. 5.20). An IIP3 of 48.5 dBm can be obtained when only RF-MEMS varactors and fixed capacitors are used at 2 GHz. The measured filter IIP3 is lower than the device specification, which is 65-67 dBm in a 50 Ω setup. The measured ACPR is > 57.8 dB for an input power of 26 dBm



Figure 5.20: (a) Measured IIP3 at 2 GHz with RF-MEMS and fixed capacitors, (b) measured IIP3 and P1dB for the 1.7 - 2.4 GHz filter with varactor diodes.

Ref.	Tun.	Ord.	I.L.	FBW	B. Vol.	IIP3	Notes	
[79]	1.45-1.89	2	2.92-2.5	66-74	1-20	6-26	Const. BW	
[80]	1.1-1.88	3	6.8-4.3	82-98	0.1-20	10.8-15.8	Const. BW	
[81]	0.8-1.35	2	3-2	38-46	2.8-22	12-18	Const. BW	
[82]	0.73-1.03	4	7-5.7	36-52	0-12	17-19.5	2-4 pole recon.	
[83]	1.25-2.1	4	6.5-3.5	54-162	0-15	13-19	Zero tuning	
[84]	1.5-2.1	4	10-4.5	40-120	0-20	5.5-17	BWC., HSBR	
This	0.7-1	4	2.9-2.6	29-80	0-20	25.2-28	BWC, HSBR	
This	1.7-2.4	4	3.6-2.6	70-178	0-20	20-31	BWC, HSBR	

 Table 5.1: Comparison with previous work

at 2 GHz (state 12) with RF-MEMS and fixed capacitors, and the ACPR is 37-43 dB for the filter with varactor diodes for an input power of 10 dBm (Fig. 5.21). Photographs of the proposed filters are shown in Fig. 5.22 and they include the filter structure, the diode DC bias lines, and the digital control lines for the RF-MEMS Cavendish Kinetics varactors.

A comparison with previous work is presented in Table 5.1. Tuning range (Tun.) is in GHz, insertion loss (I.L.) is in dB, 1-dB fractional bandwidth (FBW) is in MHz, bias voltage (B. Vol.) is in Volts, and IIP3 is in dBm. The comparison is based on tunable



Figure 5.21: (a) Measured spectral regrowth at 26 dBm for a 5-MHz WCDMA signal at 2 GHz with RF-MEMS and fixed capacitors, (b) measured ACPR versus frequency for the 1.7 - 2.4 GHz filter with varactor diodes with $P_{in} = 10$ dBm.



Figure 5.22: Photographs of the (a) 0.7 - 1 GHz filter with SPI digital control, (b) 1.7 - 2.4 GHz filter with RFFE digital control.

filters with additional advantages in terms of bandwidth control (BWC), constant bandwidth (Const. BW), reconfigurability (recon.), and high stopband rejection (HSBR). This work presented a 4-pole filter with very low insertion loss that is comparable to the insertion loss of the 2-pole filter presented in [79] and much lower than the insertion loss of the 4-pole filters presented in [83, 84]. The 1-dB FBW of the proposed filters is controllable while maintaining a low insertion loss. Furthermore, the stopband rejection is greatly enhanced by additional transmission zeroes close to the passband. The maximum capacitance ratio corresponds to RF-MEMS varactor and it is less than [79–84] at same frequencies. The IIP3 of the proposed filters are much higher than previous work and it reaches 50 dBm with RF-MEMS varactors only.

5.6 Conclusion

Two compact tunable 4-pole bandpass filters with enhanced performance are presented for applications from 0.7 to 2.4 GHz. The performance of the filters is equivalent to 4-pole bandpass filters with 4 transmission zeroes, and results in a controllable bandwidth together with a wide stopband rejection. Linearity and complex-signal measurements are done and result in a maximum IIP3 of 52 dBm and maximum ACPR of 58.3 dB. These high performance filters are suitable for applications in modern multi-standard communication systems.

5.7 Acknowledgment

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Chapter 6

Conclusion

6.1 Summary of Work

Chapter 2 presented two double bow-tie slot antennas to cover a frequency range of 100 - 300 and 200 - 600 GHz for millimeter-wave and terahertz applications. The use of a bow-tie slot significantly increases the impedance bandwidth, and a new design for the CPW low-pass filter ensures a wideband short over a 3:1 frequency range. This antenna should find applications in radio-astronomical systems or THz imaging systems required a 3:1 frequency range.

Chapter 3 demonstrated a 100 - 300 GHz quasi-optical network analyzer using compact Schottky-diode harmonic transmitter and receiver modules. The measured system EIRP and conversion gain agree well with simulations. The system is compact with a planar architecture and can be scaled to THz frequencies with the use of a smaller bow-tie antenna (300 GHz to 1200 GHz) and monolithically integrated diodes. The frequency response of two different FSS filters at 150 GHz and 200 GHz agree well with simulations. It is expected that better performance and higher dynamic range can be achieved using polarizers and a Gaussian-beam waveguide system between the transmitter and receiver units.

In chapter 4, a study of the effects of loading an antenna with set of capacitance is presented. An LTE antenna with dual feeds is presented. The antenna has the capability of tuning at multiband frequencies, thus supporting carrier aggregation and providing a wideband performance to satisfy the requirements of modern LTE standards. The design is a dual-feed quadruple-pole antenna with two tunable resonances at low-band (0.7 - 1 GHz) and two tunable resonances at high-band (1.6 - 2.6 GHz). The antenna radiation patterns are isotropic as we expected from PIFAs. The radiation efficiency is between 25-60% using varactor diodes and higher than 50% when higher quality factor tuners are used such as RF-MEMS (Q > 250 at 0.7 GHz). A new method is introduced to increase the bandwidth by adding a shunt capacitor. This new antenna is simulated inside a head and hand environment and results in a good performance making the design reliable for real applications.

In chapter 5, two compact tunable 4-pole bandpass filters with enhanced performance are presented for applications from 0.7 to 2.4 GHz. The performance of the filters is equivalent to 4-pole bandpass filters with 4 transmission zeroes, and results in a controllable bandwidth together with a wide stopband rejection. Linearity and complexsignal measurements are done and result in a maximum IIP3 of 52 dBm and maximum ACPR of 58.3 dB. These high performance filters are suitable for applications in modern multi-standard communication systems.

6.2 Future Work

For the wideband millimeter-wave and terahertz applications, dual-polarized antennas can be built to cover 3:1 frequency range and more. Antennas with differential feeding can be designed and used in different applications. The double bow-tie slot antenna can be scaled to cover frequency range > 1 THz and can be used in medical imaging systems and material characterization.

After designing the 100 - 300 GHz quasi-optical network analyzer. The system can be scaled to cover higher frequencies > 500 GHz and used for measuring different materials especially filters and for characterizing material properties. Different systems can be built to measure the magnitude and phase together. Better filters can be used to improve the performance. An integrated mono-pulse radar receiver can be similarly developed for tracking applications that required a 3:1 frequency range.

Multiple LTE antennas can be designed with multiple tuner topologies and the effects of using series and shunt tuners in small antennas need to be discussed in terms of size, linearity, and tuning range. Antennas can be built with high-Q RF-MEMS to enhance their performance in terms of radiation patterns and efficiency. Small antennas can be tested inside head and hand environment to be qualified for real applications.

Appendix A

The UCSD Fabrication Process in Details

The table on the following pages of this appendix describe the detailed fabrication process to construct the 100 - 300 GHz free-space scalar network analyzer chips as described in this dissertation. The fabrication process is done in NANO3 at University of California, San Diego.

Step	Action
1	Clean the wafer: Acetone/IPA/Methanol/DI Water
2	Photolithography:
	- Spin NR9-3000 P5 @ 4K RPM, t = 40", ACCL = 35
	- Prebake: 1' @ 150 °C
	- Expose: 10" (for lamp intensity 7.5) (Mask: Layer 1)
	- Post bake: 1' @ 100 °C
	Continued on next page

Table A.1: Fabrication Process Procedures

	- Develop in RD6:DI 3:1 for 35", rinse with DI Water
3	Sputtering:
	Sputter Ti/Au/Ti (200/3000/200 $^{\circ}$ A) Use substrate heater in
	Denton for in-situ dehydration
	- Use RF Bias for in-situ Argon plasma clean prior to sputtering
	Pressure = 4.2 mT, RF Bias = 100 W for 30"
	- Set temperature to 70 °C, wait for overshoot to 100 °C, turn off
	- DC Power = 200 W, Rotation = 65, Pressure = 4.1-4.3 mT,
	55"/6'/55"
	- Ar = 40, N_2 = 0, adjust Argon to achieve pressure
4	Lift-off:
	- Lift-off in acetone with ultrasonic: 5' (check and adjust)
	- Clean wafer with acetone/IPA/methanol/DI water
5	Inspect (microscope) and Dektak
6	Sacrificial Layer:
	PMMA-C4 (0.55um):
	- Spin @ 1700 RPM, ACCL = 255, t = 40"
	- Bake @ 135 °C for 10' then 180 °C for 2'
	- Check thickness with filmmetrics on the side of the wafer
	PMGI-SF5 (or SF6) (0.3um):
	- Spin @ 1700 RPM, ACCL = 255, t = 40"
	- Bake @ 135 °C for 10', then 180 °C for 2'
	- Check thickness with filmmetrics using PMMA program
7	E-Beam hard mask:
	Continued on next page

	- Evaporate ≈ 700 °A of Ti at 4 °A/S (use crystal for thickness)			
8	Pattern hard mask:			
	- Spin 1818 @ 4K RPM, t = 35", ACCL = 255			
	- Prebake: 90" @ 105 °C			
	- Expose: 15" (for lamp intensity 7.5) (Mask: Layer 2)			
	- Develop: MF-319 for 35"			
	- Spin NR9-3000 P5 @ 4K RPM, t = 40", ACCL = 35			
	- Prebake: 1' @ 150 °C			
	- Expose: 10" (for lamp intensity 7.5) (Mask: Layer 4)			
	- Develop in RD6:DI 3:1 for 35", rinse with DI Water			
	- Etch Ti in HF:DI water 1:10			
	- Flood expose for 45"			
	- Develop in RD6:DI 3:1 for 35", rinse with DI Water			
	- Flood expose for 45"			
	- Develop in Microdev:DI 1:1 for 30"			
9	Dry etch sacrificial layer (Oxford P80):			
	- Use O ₂ clean recipe			
	- Chiller = 35, Power = 50 W, Pressure = 50 mTorr,			
	Flow = 50 sccm, time=10'			
	- Increase power to 150 W, time = 3'			
	- Veeco			
	- Remove Ti in HF:DI water 1:10, rinse and dry			
	Continued on next page			

	- Inspect and Dektak
10	Air-bridges:
	Seed Layer:
	- Use RF Bias for in-situ Argon plasma clean prior to sputtering
	Pressure = 4.2 mT, RF Bias = 100 W for 30"
	- Set temperature to 70 °C, wait for overshoot to 100 °C, turn off
	- DC Power = 200 W, Rotation = 65, Pressure = 4.1-4.3 mT,
	40"/5'/40"
	- Ar = 40, $N_2 = 0$, adjust Argon to achieve pressure
	Mold:
	- Spin SPR-200-7
	- Prebake @ 115 °C for 1' 40"
	- Align then expose 25" hard contact, Gap 40/30, WEP: 0.3
	- Develop MF319 around 2' to 3'
	- Remove Ti by DI/HF: 9:1
11	Gold electroplating:
	- Add water up to 800-1200
	- Temp: 65 °C, Stir: 150-200
	- Connect +ve to Anode and ve to sample
	- Start with a current of 6 mA for 10-15 min and check
	a. If color is black, increase the current
	b. If it is rough, decrease the current
12	Remove mold:
	- Flood Expose: 1'
	Continued on next page

	- Develop: Microdev	
13	Etch the seed layer:	
	- Spin 1818 like sacrificial layer	
	- Etch Ti/Au/Ti	
14	CPD (Tousimis) and release:	
	- Clean machine, fill with methanol	
	- Set purge time to 15'	
	- Use settings written on machine	

Appendix B

A Tunable Dual-Band Single-Feed Triple-Resonance LTE Antenna for Carrier Aggregation Systems

B.1 Introduction

With the rapid growth of wireless communications, high data-rates are becoming attractive in modern systems. This will require a new generation of small wideband antennas. The idea of carrier aggregation (CA) is used in LTE-Advanced in order to increase the bandwidth, and therefore increase the data rate. In this new standard, the transmit or receive (or both) data is divided between different carriers spaced 100's of MHz apart so as to improve the data rate and improve the system efficiency.

In this chapter, a small PIFA antenna is designed to satisfy the CA condition by resonating the antenna at three different frequencies. In this case, the antenna can be used either in a standard configuration with one or two resonances tuned to the transmit frequency and one or two resonances tuned to the receive frequency, or one can tune the



Figure B.1: Antenna layout: (a) top view, (b) side view.

resonances to the three different carries for CA applications.

B.2 Antenna Design

The antenna layout is shown in Fig. B.1, with a size of 36×16 mm. The antenna is integrated on $66 \times 100 \times 3.154$ mm circuit board using 2-ply 60 mils Rogers RO4003C ($\varepsilon_r = 3.55$, $tan\delta = 0.0027$ at 2.5 GHz) stacked together using 1-ply 4 mils Rogers RO4450B ($\varepsilon_r = 3.54$, $tan\delta = 0.003$ at 10 GHz), the thickness is enough for 0.7 - 2.7 GHz applications. The location of the varactor diodes is chosen using Ansys-HFSS simulator [31], and divides the current distribution into three arms causing the antenna to resonate at three different frequencies. Diode 1 forms the longest path between the open and short circuits causing the first resonant frequency, while the diode 2 forms the second longer path causing the second resonant frequency. Diode 3 forms the shorter path causing the higher resonant frequency. The feed is located at the center to provide a wideband impedance match over the entire frequency range.

B.3 Simulated and Experimental Results

B.3.1 Input Matching

The antenna is placed on a 10×6.6 cm printed-circuit board with a continuous ground plane on the back (antenna is placed at the top corner), and the simulated results were obtained using Ansoft-HFSS. The antenna shows a wideband impedance match from 0.7 - 1 GHz (Fig. B.2a). The first resonant frequency can be tuned from 0.7 - 0.85 GHz. The second resonant frequency can be tuned from 0.85 - 1 GHz. The required capacitance for the first resonant frequency is 3-12 pF, while the second resonant frequency needs a capacitance range of 2-5 pF. The third resonant frequency can be tuned from 1.8 - 2.5 GHz (Fig. B.2b) and needs a capacitance of 0.63-2.67pF. Both diodes are commercial units obtained from Skyworks with a control voltage of 1-30 V. Measurements agree well with simulations (not shown). A shunt capacitor (C_m) is used to control the harmonics of the low-band resonances and not interfere with the high-band resonance.

B.3.2 Radiation Patterns and Efficiency

The measured radiation patterns are shown in Fig. B.3, and are nearly isotropic, as expected. The measured and simulated antenna efficiency is presented in Fig. B.4. The simulation results do not include the mismatch loss and therefore, are correct at the center frequency of the tuned band. Simulations are done with tuners of different quality factors (Q). The antenna efficiency is 25-50% at lower band and 50-80% at higher band. The efficiency will increase to 50-90% using RF MEMS (Q > 250) based on simulations. In general, Schottky diodes are low Q (40-80) devices and are only used here for demonstration.



Figure B.2: Measured reflection coefficient (S_{11}) versus frequency at (a) low-band, (b) high-band.



Figure B.3: Measured antenna patterns at 0.9 and 2.2 GHz.



Figure B.4: Measured and simulated radiation efficiency of the tunable antenna.



Figure B.5: A picture of the 4G-LTE Antenna on PCB.

B.4 Conclusion

In this chapter, a tunable 4G-LTE antenna with a single feed that resonates at three different frequencies has been demonstrated. Two resonances are tunable within the LTE low band (0.7 - 1 GHz) and one resonance is tunable within the LTE high band (1.7 - 2.5 GHz). The antenna has nearly isotropic patterns and radiation efficiency of 25-80% using Schottky diodes.

Appendix C

A Low-Loss 1.4-2.1 GHz Compact Tunable Three-Pole Filter With Improved Stopband Rejection Using RF-MEMS Capacitors

C.1 Introduction

RF-MEMS technology is experiencing a substantial development effort to enhance its capabilities and commercial vendors are now producing MEMS packaged devices with high reliability. This include Cavendish Kinetics and wiSpry for capacitive switches [70, 92], and other manufacturers are working on high-reliability ohmic contact switches [93,94]. These devices are small ($\approx 2 \text{ mm}^2$ bumped die), hermetically-packaged, and ESD protected with excellent long-term reliability under medium to high RF power conditions (up to 2 W). Cavendish Kinetics switched capacitors, for example, can handle RF power of 33 dBm with cycling > 10B cycles and switching time



Figure C.1: (a) Circuit model of the tunable 3-pole filter structure, and (b) equivalent circuit of the coupled lines.

 $< 50 \ \mu$ s. They are also easy to control with SPI or RFFE digital controls. Several designs were recently demonstrated using these devices, in particular, RF filters and phase shifters [95, 96]. The results of [95, 96] show a substantial improvement in terms of insertion loss and linearity.

Three and four-pole tunable filters with extra transmission zeroes were demonstrated in [83, 85]. These zeroes are located close to the passband to improve the filter stopband rejection and enhance the filter selectivity. This chapter presents a compact tunable 3-pole bandpass filter with low-loss and improved stopband rejection using RF-MEMS capacitors and varactor diodes.

C.2 Filter Design

The proposed tunable 3-pole band-pass filter is shown in Fig. C.1(a). This filter consists of two $\lambda_d/4$ combline resonators loaded by RF-MEMS Cavendish Kinetics varactors, DVC₁-DVC₂. The opposite ends with via holes are modeled using smallvalue inductors instead of ideal short circuits. The via-hole inductance determines the filter bandwidth if the filter rejection is required to be the same. For the same rejection, the filter with a larger via-hole inductance (narrower diameter or thicker substrate) has a wider bandwidth. The input and output coupling is achieved using a tapped-line connection, and series varactors C_{IN} are used to obtain a good impedance match over a wide frequency range. The coupling values between the adjacent resonators include the intrinsic coupling of the coupled lines, which are equivalent to the circuit shown in Fig. C.1(b), and the capacitive coupling introduced by the varactor diode C_{12} . For the non-adjacent resonator, there is an additional coupling path consisting of transmission lines and varactor diode C_{13} for generating extra pole and transmission zero.

The simulated (S_{21}) with and without the non-adjacent resonator is shown in Fig. C.2. The two filters are optimized to have the same 1-dB FBW for a fair comparison. The coupling path can be also seen as a resonator (a back-to-back combline resonator) which is intentionally designed to have a resonant frequency in the passband. As a result, a 3-pole filter response is obtained with three transmission zeroes to improve the stopband rejection.

The resonant frequency of a combline-line resonator is determined by the characteristic impedance and the length of the $\lambda/4$ transmission line in addition to the loading capacitor. If the electrical length is short and less than 30° at ω_0 , the resonant frequency can be approximately calculated by [99]

$$\omega_{0i} \approx \sqrt{\frac{\omega_0}{Z_i \theta_i C_i}}, \quad i = 1, 2$$
 (C.1)

$$\omega_{013} \approx \sqrt{\frac{\omega_0}{2Z_{13}\theta_{13}C_{13}}} \tag{C.2}$$

The external Q is evaluated by

$$Q_e \approx \frac{R_0 \omega_0}{\omega_{01} Z_1 \theta_T} \approx \frac{R_0}{\theta_T} \sqrt{\frac{\omega_0 C_1 \theta_1}{Z_1}}$$
(C.3)

The equations of the transmission zeroes which are obtained using even- and odd-mode



Figure C.2: Filter response with and without the non-adjacent coupling path.

analysis are complicated and are not addressed in this chapter due to brevity.

A tunable 3-pole band-pass filter is designed to cover a frequency range of 1.4 to 2.1 GHz. The filter layout is shown in Fig. C.4 with a size of $9 \times 5 \text{ mm}^2$. The filter is printed on 50 mils Rogers RT/duroid 6010LM board ($\varepsilon_r = 10.7$, $tan\delta = 0.0023$ at 10 GHz) and soldermask covers the transmission-lines to reduce the loss introduced by the electroless nickel immersion gold (EPIG) metal finish [95]. The length of the combline resonators are 3.1 mm (15.7° at $f_{0,max}$) and allow the use of equations (C.1 and C.2) to determine the filter operating frequency with a resonator impedance of $\approx 50 \Omega$. Three hole-vias with a radius of 0.2 mm each are located between the two resonators so as to provide a ground connection. Large vias are employed for a smaller inductance and a narrower bandwidth with the same rejection.

The loaded $\lambda/4$ resonators are tuned using Cavendish Kinetics CK505S capacitor. The 5-bit Cavendish Kinetics RF-MEMS Digital Variable Capacitor (DVC) is built on a 0.18 μ m CMOS platform with chip-scale hermetic packaging and flip chip bumps, as shown in Fig. 5.4(a) [70]. Its overall size is 1.4×1.3 mm², and includes the RF-MEMS capacitor banks (at the bottom) and the control circuits and charge pump (at the top). The device requires 1.8 V to operate and is easily integrated into existing communication systems. A capacitance value of 1.1-5.1 pF is achieved over 32 states with a capacitance step of 125 fF. The device is controlled using a serial peripheral interface



Figure C.3: CK device with SPI control: (a) block diagram, (b) capacitance and quality factor at 1.75 GHz.

(SPI) where three digital lines (CLK, SDA, and CSN) are required. The measured device Q is 253-48 at 1.75 GHz from state 0 (C = 1.1 pF) to state 31 (C = 5.1 pF) and Q of 63 is obtained at state 21 (C = 3.8 pF) as shown in Fig. 5.4(b). The measured IP3 is 65-67 dBm in a 50 Ω set-up over the entire capacitance range.

The coupling capacitor between the resonators (C₁₂) is tuned using back-to-back series Skyworks diodes SMV-2020-079LF, with $C_{12} = 0.35 - 3.2$ pF, $R_s = 2.5 \Omega$, L = 0.7 nH, and $V_{Bias} = 0.20$ V. The auxiliary resonator (C₁₃) is tunable using a Skyworks diode SMV-1413-079LF, with $C_{13} = 1.77 - 9.24$ pF, $R_s = 0.35 \Omega$, L = 0.7 nH, and V_{Bias} = 0-30 V. A 2.7 pF high-Q ATC capacitor is connected in series with the C₁₃ diode for DC bias and also to control the capacitance range. Two C_{IN} varactor diodes are required for wideband impedance matching, and are connected to the filter input and output ports and placed in series with 27 pF high-Q ATC capacitors for biasing. The C_{IN} diodes are Skyworks diodes SMV-1235-079LF, with C_{IN} = 2.38 - 18.22 pF, $R_s = 0.6 \Omega$, L = 0.7 nH, and $V_{Bias} = 0.15$ V. Resistors of 30 k Ω are used to prevent any RF leakage to the



Figure C.4: Layout of the compact 1.4 - 2.1 GHz tunable 3-pole band-pass filter. DC bias lines.

C.3 Measurements

Fig. C.5 presents the measured filter S-parameters (S_{21} , S_{11}). The measured filter tuning range is 1.4 to 2.1 GHz and was obtained using 21 capacitance steps (Fig. C.6a). Note that C_{12} and C_{13} are tuned accordingly at each DVC₁ and DVC₂ capacitance step. The measurement is stopped at state 21 due to the capacitance range limitation of C_{12} and C_{13} . The DC voltages required to tune the C_{12} and C_{13} diode are 0-13.5 V and 0-24 V, respectively. A DC voltage of 6-15 V is also used to tune C_{IN} for impedance matching from 1.4 to 2.1 GHz. Note that a tuning range down to 1.15 GHz can be obtained if different C_{12} and C_{13} diodes are used (Fig. C.6a).

The filter insertion loss is 1.73-2.0 dB with a 1-dB FBW of 12-13% (160-250 MHz). The measured S_{11} is < -20 dB for all states. Two transmission zeroes are presented at both sides of the passband and provide a wideband stopband rejection > 30 dB. Measurements agree well with simulations done by Sonnet and ADS [32, 98] (not shown for brevity).

The shape factor of the filter is calculated from the measured S_{21} as the ratio $(f_4 - f_1)/(f_3 - f_2)$ where the insertion loss is 6 dB at f_2 and f_3 and 30 dB at f_1 and f_4 .



Figure C.5: Measured 1.4 - 2.1 GHz S-parameters: (a) S_{21} , (b) S_{11} . Only 9 states are shown out of 21 states.

The lower the shape factor, the steeper the skirts of the filter's response curve. Ideally, the shape factor equals 1 (rectangular shape). A measured shape factor of 1.75-1.8 is obtained at 1.4 - 2.1 GHz.

C.4 Conclusion

A compact 3-pole tunable filter with enhanced performance is presented. The filter performance is equivalent to a 3-pole filter with 3 zeroes, and results in a low insertion loss with a wide stopband. The low loss is due to the high-Q RF-MEMS tunable capacitors used.



Figure C.6: Compact 1.4 - 2.1 GHz filter: (a) center frequency vs. DVC state, (b) measured 1-dB bandwidth and loss vs. center frequency.



Figure C.7: A photograph of the 1.4 - 2.1 GHz filter with SPI digital control.

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