EXPERIMENTAL INVESTIGATION OF NEW INDUCTOR TOPOLOGIES

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ABSTRACT

EXPERIMENTAL INVESTIGATION OF NEW INDUCTOR TOPOLOGIES

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Inductors have been investigated and used extensively in analog circuits and digital circuits over 200 years. This dissertation presents a 3D inductor which can efficiently work with a high Q variable capacitor (varactor), and two types of tunable inductors using vanadium dioxide (VO$_2$) as phase control material. There are three different 3D inductor models that are designed and verified using electromagnetic simulation tools, and applied in a resonator with a voltage tunable ferroelectric thin-film barium strontium titanate (BST) varactor. The resonator combining the 3D inductors and BST thin film varactor is verified for the first time.

Two types of tunable inductors are designed using VO$_2$ thin films which exhibit metal-to-insulator (MIT) phase transition characteristics above a transition temperature. A VO$_2$-based integrated actuator, which has two-way displacements, is designed and used in the first tunable inductor design. The integrated actuator moves a Nickel layer, which has a high permeability (~100), over a planar coil inductor to tune the overall inductance. Simulation results show a 27% tuning range through applied displacement variation.
(5μm - ∞) in the simulator. The second tunable inductor design uses a VO₂ thin films as series switches in the devices. One configuration uses a short VO₂ bar as a circuit switch and the other uses VO₂ to replace the full spiral coil inductor. The total inductance of the circuit tunes when the VO₂ changes from insulating to conducting phase above the transition temperature. The experimental results show 32% inductance variation through applied temperature variation (25°C - 100°C) in the devices. The simplified fabrication process of the second tunable inductor exhibits potential for portable and adaptive communication systems. Two configurations of multi-tap inductors are also designed such that the inductance of the device changes when the probe is moved between the output ports. A strong match between the VO₂ based devices and the Au based devices reveal that we are able to produce high quality VO₂ thin films. The VO₂ thin film has never been utilized in electrical application such as inductors before this research.
Dedicated to my parents, my wife, and my younger sister.
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CHAPTER I
INTRODUCTION

1.1. Motivation

The study of inductor started in 1830 when Michael Faraday proposed the first model. Since then, the inductors have existed in many applications from power conditioning circuits to RF and microwave circuits. However, the traditional inductors inherent performance, integration issues and relatively large size restrict their utilization in modern circuits. In the study of inductors design and applications, lots of novel structures such as 3D inductor and tunable inductor have proved to effectively improve the circuit and system performance at RF and microwave frequencies because of 3D inductor having improved quality factor, and tunable inductor achieving the same capability that multiple conventional inductors are normally required for.

Several research groups successfully designed and fabricated high performance small 3D inductors with different configurations [1]–[4]. Rao [1] presented a new 3D inductor structure that improved the inductance ~ 19% - 23% higher than the traditional planar coil inductor. The Q factor achieved was 23 - 25 at microwave and higher frequencies range (30 – 100GHz). Tang [2] fabricated multi-layer 3D inductors by standard one-poly-four-metal (1P4M) CMOS process. These 3D inductors only occupied 16% of the area of
the conventional planar inductor when the inductance values are the same. Radosavljevic [3] designed, fabricated, and measured a huge 3D inductor by Low Temperate Co-fired Ceramic (LTCC) technology. The Q factor achieve to 72 at 1 GHz with a very large device seize (8mm×5mm). Weon [4] presented a very high Q factor 3D inductor by stressed metal technology. The Q factor of inductors on a low-resistivity silicon substrate is 75 at 4 GHz when the inductance is 1nH. The Q factor of inductors on a high-resistivity silicon substrate is very impressive to 140 at 12 GHz when the inductance is 1.2nH.

The tunable inductors have flexible design and stable performance in microwave and RF frequency. Stable high quality tunable inductors could be widely used in communication systems for tunable filter components, impedance matching network, voltage-controlled oscillators, amplifiers, and phase shifter circuits. Four popular approaches are used currently for designing a tunable inductor are: 1) utilizing switches or relays to separate or group inductor segments; 2) controlling magnetic materials in the tunable inductors to tune the magnetic field in the inductors; 3) controlling the magnetic coupling coefficient to tune the mutual-inductance between the primary coil and secondary coil; 4) moving metal structure over the inductor coil to tune the magnetic flux through the inductors. Each of the approaches are widely used for designing high performance tunable inductors.

Figure 1.1(a) shows the schematic model of the tunable inductors using micro-switches and micro-relays. The switches or relays are used to control the inductance directly (self-inductance, Figure 1.1(a)) [5], [6] or indirectly (mutual-inductance, Figure 1.1(b)) [7]. Both the initial inductance are L1 when the switches are open. Switch closing
will increase the total inductance to \( L_1 + L_2 + L_3 \) in case (a), and change the equivalent inductance to \( L_{eq} = L_1 + L_{12} + L_{23} \) in case (b).

![Schematic model of the tunable inductor with switches and relays. (a) Inductors are tuned directly. (b) Inductors are tuned by changing mutual-inductance.](image)

Figure 1.1  Schematic model of the tunable inductor with switches and relays. (a) Inductors are tuned directly. (b) Inductors are tuned by changing mutual-inductance.

The second approach for a tunable inductor is to change the effective permeability of the magnetic materials in the inductor. Tunable permeability materials could be tuned by external stimuli such as DC bias, high temperature, or low temperature [8]–[10]. These materials was usually be used in coil inductors, as shown in Figure 1.2(a). Another popular approach recently is to move the magnetic materials position [11] that will change the effective permeability of the coil, as shown in Figure 1.2(b). These inductors could be also around by some magnetic materials such as salt water [12] and liquid ferromagnetic materials [13]. The problem in this approach is the complex fabrication process because the specific magnetic materials. There is not a general process to manufacture these inductors. However, the various processes provide flexible designs and structures.
Figure 1.2 The structures of tunable inductor with tunable magnetic materials. (a) Apply DC bias to tune the permeability of the magnetic materials. (b) Move magnetic materials position to tune the effective permeability.

Controlling coupling coefficient between the primary coil and secondary coil to result the mutual-inductance is the third way to design tunable inductors [14]–[19]. The coupling coefficient can be written as

\[ k = \frac{L_M}{\sqrt{L_p L_s}} \]  

(1)

where \( L_p \) and \( L_s \) are the inductance of primary coil and secondary coil. The \( k \) is the coupling coefficient. The \( L_M \) is the mutual-inductance of the tunable inductor. Figure 1.3 is the equivalent circuit model of the coupling tunable inductor.

Figure 1.3 Schematic model of the coupling tunable inductors.
Utilizing removable metal plate is the last approach of the tunable inductors. The inductor is tuned by a moving metal plate [20]–[22]. The metal plate shield the magnetic flux of the inductor, thereby the inductance is changed by the metal location when the metal plate is moving, as shown in Figure 1.4. The main difference between this approach and the coupled inductor is that this approach doesn't have to use magnetic materials. However, it has to move the metal plate always.

Figure 1.4 Structures of tunable inductors with removable metal plate.

1.2. Objectives

As a continuing effort in the search for 3D inductor, there are two objectives in the 3D inductor study: 1) Design and understand three 3D inductor models using electromagnetic simulation tools. 2) Develop a resonator based on the 3D inductors and an existing high quality variable capacitor (varactor) [23] with barium strontium titanate (BST) thin film. We believe that the existing conventional planar inductors do not have sufficient Q factor to provide the performance needed by the varactor based circuits. The
3D inductor has proved an effective structure to improve the performance of inductor further. Such circuits can then be utilized in a variety of electronic applications such as electronically tunable filters which can be widely used in cell phones and other communication systems.

There are three objectives in the tunable inductor study: 1) design a tunable inductor using vanadium dioxide (VO$_2$) as cantilever. A VO$_2$-based actuator is utilized to move a Nickel pad up and down over a conventional coil inductor. The inductance is changed by the varied effective permeability in the inductors when the Nickel is moved due to the movement of VO$_2$ and Au cantilevers above the transition temperature. 2) Design two configurations of tunable inductors using VO$_2$ as switches. The inductance change when the VO$_2$ thin film change to conducting phase above the transition temperature. 3) Design two configurations of multi-tap inductors using VO$_2$ as switches. The overall inductance changes when the probe is moved between the output ports.

The new actuator in the first design improves the performance of the existing actuator which can only provide one direction (upward) movement [24]. This new research utilizes VO$_2$ as an actuator during phase transition, which have never been utilized in electrical application such as inductors. The tuning range will also be much wider than the tunable inductors in the literature [15], where only bimorph cantilevers (downward movement only) are used to move a metal loop. The second tunable inductor design and the multi-tap inductor design provide longer life time, smaller device size, and lower manufacturing cost than the existing similar tunable devices due to the simplified fabrication of VO$_2$ thin film.
1.3. Contributions

This dissertation focuses on new designs of 3D inductors and tunable inductors. A resonator combining the 3D inductors and BST thin film varactor is also verified for the first time. A complex fabrication process is developed to create an effective approach to utilize the 3D inductors in electrical applications such as resonators and filters. Before this research, the VO$_2$ thin film has never been utilized in electrical application such as inductors. Two tunable inductors with different topologies are designed using VO$_2$ actuation and metal-to-insulator characteristics respectively. The new actuator in the first design improves the performance of the existing actuator. The second tunable inductor design leads to a much simplified fabrication process that would have good benefits such as overall system versatility, size reduction and decreased cost.

Here are the publications that resulted from this dissertation work:

1.4. Outline

This dissertation starts with a study of 3D inductor models and developed resonators based on the 3D inductors and high quality varactor in chapter II. In chapter III, two types of tunable inductors and a multi-tap inductor model using VO\textsubscript{2} thin film are presented. This dissertation ends with the conclusions and future work in chapter IV.
CHAPTER II

3D INDUCTOR AND HIGH K DIELECTRIC THIN FILM BASED RESONATOR CIRCUIT

This chapter starts with a review of the BST thin film and the high quality varactor in section 2.1. In section 2.2, three 3D inductor models are presented and studied based on electromagnetic theory. The method of the fabrication for the 3D inductor and developed resonator is introduced in section 2.3. In section 2.4, the simulated and experimental results are analyzed in terms of the electromagnetic model and equivalent circuit.

2.1. Varactor Review

2.1.1. Barium Strontium Titanate (BST)

Ba$_x$Sr$_{1-x}$TiO$_3$ have been investigated [25]–[29] for the reconfigurable RF and microwave components and circuits for many years because of its good tuning ratio at room temperature, stable and low loss-tangent, and low power consumption. The dielectric constant of BST thin film is tunable using both negative and positive voltage due to its varied dielectric strength when a voltage bias is used on it. Ba$_{0.6}$Sr$_{0.4}$TiO$_3$, is composited using barium tinanate (BaTiO$_3$) and strontium titanate (SrTiO$_3$), have been used in our lab to fabricate high quality components for many years [23], [30]–[33]. Its
dielectric constant decrease continually from 700 to 150 when the DC bias voltage increase from 0V to 8V. The large tuning ratio ($\sim4:1$) of the BST thin film at room temperature exhibit potential for high quality tunable applications.

2.1.2. Pulsed Laser Deposition (PLD) System

A Pulsed Laser Deposition (PLD) process have been utilized by our group to obtain high quality BST thin film on sapphire and silicon substrates. PLD is a physical vapor deposition technique. The material is vaporized from the target into plasma plume using a high energy laser beam to strike continually on the target in a high vacuum chamber before it is deposited on a heated substrate. In our large area PLD system, as shown in Figure 2.1, a rotary pump and a high vacuum turbo pump are used to evacuate the chamber to avoid the contaminants in the chamber. The chamber oxygen background pressure is maintained at 75 mT during the deposition process. A rotated target is struck by a high-power pulsed laser beam in a high vacuum chamber (typically 175mJ – 180mJ energy with 30Hz frequency). Then a plasma plume is created and ejected from the target to form the plume in the chamber. Then the dynamic plasma deposit and grow on a rotated heated substrate (900°C). The thickness of the thin film is controlled by the laser repetition rate, and the thin film quality is controlled by the laser parameters, chamber window transmission, chamber temperature, and the oxygen pressure.
Figure 2.1 Pulsed Laser Deposition System. (a) Large are PLD system in our lab. (b) PLD simplified setup.
2.1.3. Thin-film Varactor based on BST

A high quality varactor, is similar to the method proposed in [23], have been considered to be applied in this project. The configuration of the varactor is shown in Figure 2.2 where the varactor is based on an 180nm BST tunable layer sandwiched between two metal layers. The deposition method for the BST thin film is based on the Pulsed Laser Deposition (PLD) System. An equivalent schematic circuit, as shown in Figure 2.3, is used to match the experimental results in order to find out the capacitance value of the varactor. Figure 2.4 present the experimental and simulated insertion loss and return loss versus varied DC bias voltage. The decreased insertion loss using higher bias voltage reveals that more signal pass through the varactor to the output port due to its decreasing capacitance. The capacitor values are confirmed by the schematic model and plotted in Figure 2.5. The dielectric constant of the BST thin film is confirmed by the equation

\[ \varepsilon_r = \frac{C \times d}{A \times \varepsilon_0} \]  \hspace{1cm} (1)

where the \( \varepsilon_0 \) is the vacuum permittivity \((8.854 \times 10^{-12} F \cdot m^{-1})\), \( A \) is the overlap area of the varactor \((5\mu m \times 5\mu m)\), and \( d \) is the thickness of the BST thin film \((180nm)\). The varied dielectric constant value is presented in Figure 2.6(a) where approximately 4:1 tuning range from \( \varepsilon_r \) of 750 at zero bias to 180 at the highest bias voltage is obtained. The dielectric quality factor of the varactors in Figure 2.6(b) confirmed by the equation

\[ Q = 2\pi f \times C \times R_C \]  \hspace{1cm} (2)

The Q factor increases with increasing frequency or increasing effective shunt resistance of the capacitor \( (R_C) \).
Figure 2.2  Configuration of a $5\mu m \times 5\mu m$ varactor. (a) Top view. (b) 3D view.
Figure 2.3  The equivalent circuit model of the varactor.

Figure 2.4  Experimental (a) insertion loss and (b) return loss for a varactor with 5μm×5μm overlap area and 180nm BST thin film (Blue dashed line is the simulation).
2.2. **Electromagnetic Theory and Methods**

This section will address and learn the 3D inductor using the basic electromagnetic theory which will be applied in the simplest case of the self-inductance of a single wire at
the beginning. Then the theory is extended to the mutual-inductance between parallel wires and non-parallel wires cases. The inductance of 3D inductor models will be figured out and calculated in the last part of this section.

2.2.1. Single Straight Wire

A single straight wire with the self-inductance is given by the general formula in [34]

\[ L = \frac{\mu_o \mu_r}{2\pi} l [\ln \frac{2l}{r} - 1 + \frac{b_1}{l}] \] (3)

where \( \mu_0 \) is the magnetic permeability in vacuum that is \( 4\pi \times 10^{-7} H \cdot m^{-1} \), \( \mu_r \) is the relative permeability of the material around the wires, the \( r \) is the geometric mean distance of the wires, the \( \delta_1 \) is the arithmetic mean distance of the points of the cross section, and the \( l \) is the length of the wire. Round cross section wires and rectangular cross section wires are the two most general geometric configuration, are given by

\[ L_{\text{round}} = \frac{\mu_o \mu_r}{2\pi} l [\ln \frac{2l}{\rho} - 1 + \frac{\mu_r}{4}] \] (4a)

\[ L_{\text{rectangular}} = \frac{\mu_o \mu_r}{2\pi} l [\ln \frac{2l}{B + C} + \frac{1}{2} - \ln e] \] (4b)

where \( \rho \) is the radius of the round wires, \( B \) and \( C \) are the length and width of the cross section of rectangular wires.

2.2.2. Parallel Wires

*Mutual-inductor of two equal paralleled rectangular wires*

The top view of two paralleled equal wires is shown in Figure 2.7. The center-to-center distance between the two wires is assumed to \( d \), and the length of each wire is assumed to be \( l \), the exact formula for the mutual-inductance is expressed by [34]
\[
M_{\frac{a}{d}<0.1} = \frac{\mu_0 \mu_r}{2\pi} \frac{1}{2} \frac{l^2}{d} \left[ 1 - \frac{1}{12} \frac{l^2}{d^2} + \frac{1}{40} \frac{l^4}{d^4} \right]
\] (5a)

\[
M_{0.1<\frac{l}{d}<10} = \frac{\mu_0 \mu_r}{2\pi} l \ln \left( \frac{l}{d} + \frac{1 + \frac{l^2}{d^2}}{\sqrt{1 + \frac{l^2}{d^2} + \frac{d}{l}}} \right) - \frac{1 + \frac{l^2}{d^2} + \frac{d}{l}}{4}
\] (5b)

\[
M_{\frac{a}{d}>10} = \frac{\mu_0 \mu_r}{2\pi} l \ln \left( \frac{2l}{d} - 1 + \frac{l}{d} - \frac{1}{4} \frac{l^2}{d^2} + \cdots \right)
\] (5c)

Figure 2.7 Mutual-inductance between two equal paralleled rectangular wires.

**Mutual-inductance of two unequal paralleled rectangular wires**

The geometric configuration of two unequal paralleled wires is shown in Figure 2.8, the general formula for the mutual-inductance of the wires is written as [34]

\[
M = \frac{\mu_0 \mu_r}{4\pi} \left[ \alpha \sinh^{-1} \frac{\alpha}{d} - \beta \sinh^{-1} \frac{\beta}{d} - \gamma \sinh^{-1} \frac{\gamma}{d} + \delta \sinh^{-1} \frac{\delta}{d} \right.
\]

\[
-\sqrt{\alpha^2 + d^2} + \sqrt{\beta^2 + d^2} + \sqrt{\gamma^2 + d^2} - \sqrt{\delta^2 + d^2}
\]

(6)

where \(\alpha = l + m + \delta, \beta = l + \delta, \gamma = m + \delta\). The \(\delta\) is a negative number if the wires are overlapped, else \(\delta\) is positive.
2.2.3. Non-parallel Wires

The mutual-inductance of any non-parallel wires in 3D inductor model could be calculated by the model of two straight filaments placed in any desired position [34]. The geometric configuration is shown in Fig. 1.9, the general formula for the mutual-inductance of the wires is written as

\[
M = \frac{\mu_0 \mu_r}{2\pi} \cos \epsilon \left\{ [(\mu + l) \tanh^{-1} \frac{m}{R_1 + R_2} + (\nu + m) \tanh^{-1} \frac{l}{R_1 + R_4} - \mu \tanh^{-1} \frac{m}{R_3 + R_4} - \nu \tanh^{-1} \frac{l}{R_2 + R_3}] - \frac{\Omega d}{\sin \epsilon} \right\}
\]

(7)

in which

\[
\Omega = \tan^{-1} \left[ \frac{d^2 \cos \epsilon + (\mu + l)(\nu + m) \sin^2 \epsilon}{dR_1 \sin \epsilon} \right]
\]

\[
- \tan^{-1} \left[ \frac{d^2 \cos \epsilon + (\mu + l)\nu \sin^2 \epsilon}{dR_2 \sin \epsilon} \right]
\]

+ \tan^{-1} \left[ \frac{d^2 \cos \epsilon + \mu \nu \sin^2 \epsilon}{dR_3 \sin \epsilon} \right]
\]

(8)

Figure 2.8 Mutual-inductance between two unequal parallel rectangular wires.
\[
-\tan^{-1}\left[\frac{d^2 \cos \epsilon + \mu(v + m) \sin^2 \epsilon}{dR_4 \sin \epsilon}\right]
\]

and the distance

\[
R_1 = d^2 + (\mu + l)^2 + (v + m)^2 - 2(\mu + l)(v + m) \cos \epsilon \quad (9a)
\]
\[
R_2 = d^2 + (\mu + l)^2 + v^2 - 2(\mu + l)v \cos \epsilon \quad (9b)
\]
\[
R_3 = d^2 + \mu^2 + v^2 - 2\mu v \cos \epsilon \quad (9c)
\]
\[
R_4 = d^2 + \mu^2 + (v + m)^2 - 2\mu(v + m) \cos \epsilon \quad (9d)
\]

Figure 2.9 Mutual-inductance between two non-paralleled rectangular wires.

2.2.4. Methods for 3D Inductor

Three 3D inductor models are designed in this project and shown in Figure 2.10 where the wires are 10 \(\mu m\) wide and have 10 \(\mu m\) high vias between the top and bottom wires. The coils are increased from 1 turn to 3 turns. The square and circular models have similar configuration and electromagnetic fields. Therefore, only the zigzag model and the square model will be analyzed in this section using the basic electromagnetic theory. All the three models will be simulated, fabricated, and measured in the next sections. The 3D inductors can be utilized in a resonator with a 5 \(\mu m\) \(\times\) 5 \(\mu m\) varactor [23] base on a 250nm barium
strontium titanate thin film. When a DC bias is applied on the resonator, the resonance points should move to higher frequency due to the decreased capacitance on the varactor. Such circuits can then be utilized in a variety of electronic applications such as high Q filters which can be widely used in communication systems.

Figure 2.10 Top view of 3D inductor. a) 3D inductor with 3 zigzag coils. b) 3D inductor with 3 square coils. c) 3D inductor with 3 circular coils.
Zigzag Model

Figure 2.11 is the 3D view of the triangular model which include the parallel wires, non-parallel wires, and straight rectangular wire. All the posts are considered to be vertical parallel wires, as shown in Figure 2.12. The mutual-inductance of two equal paralleled wires are learned from equation (5) in the last section. The total mutual-inductance is given by the following equation:

\[ M_{\text{vertical}} = \sum_{i=1}^{N-1} \sum_{j=i+1}^{N} (-1)^k M_{(i,j)} \]  \hspace{1cm} (10)

where \( M_{(i,j)} \) is the mutual-inductance between any two posts wires, and correspond to equation (5). \( i \) is the number of the first segmented wire in the group, and \( j \) is the number of the second one. The variable \( k \) is 1 if the sum of \( i \) and \( j \) is odd due to the opposite current, else \( k \) is 0. The length of wire \( l \) in equation (5) is expressed by the height of the post wires, and the space \( d \) is expressed by the distance between the two post wires. The distance \( d \) could be written as

\[ d = \sqrt{(k' \times W)^2 + (k \times L)^2} \]  \hspace{1cm} (11)

where \( k' = (j - i)/2 \), \( W \) is the horizontal width of each turn, and \( L \) is the vertical length of the coil.

![Figure 2.11 3D view of the triangular model.](image)

21
Figure 2.12  Geometric configuration of the posts.

The mutual-inductance between the horizontal wires, which are on the same layer (top or bottom layer), belongs to the unequal parallel case in the last section. The currents flow pass through these wires with same direction, so there is no negative mutual-inductance between these wires. Figure 2.13 present the top mask layer geometric configuration of a 3 turns coil. The total mutual-inductance can be written as

\[
M_{\text{horizontal(paralleled)}} = \sum_{i=1}^{N-1} \sum_{j=i+1}^{N} k'' M_{\text{paralleled}(i,j)}
\]

(12)

where \( M_{\text{paralleled}(i,j)} \) is the mutual-inductance between any two horizontal wires are on the same layer, and correspond to equation (6). The odd \( i \) represent the wires on the top layer, and even \( i \) represent the bottom wires. The variable \( k'' \) is 1 if the sum of \( i \) and \( j \) is even, else \( k'' \) is 0. The \( \delta \) in equation (6) could be written as

\[
\delta = -(l - k' \times W \times \cos \phi)
\]

(13)

where \( \phi \) is the angle between these wires and the horizontal axis, and \( l \) is the length of each wires. Additionally, the distance between two wires \( d' \) is written as
\[ d' = k' \times W \times \sin \phi \]  

Figure 2.13  Geometric configuration of paralleled wires on the same layers.

The mutual-inductance between horizontal wires, which are on two different layers (top or bottom layer), belongs to non-parallel case in the last section. Figure 2.14 exhibit the geometric configuration of the first top wires with other bottom segments. The total mutual-inductance could be written as

\[ M_{\text{horizontal(non-parallel)}} = \sum_{i=1}^{N-1} \sum_{j=i+1}^{N} k \times M_{\text{non-parallel}}(i,j) \]  

where the \( M_{\text{non-parallel}}(i,j) \) is the mutual-inductance between any two non-parallel wires are on different layers, and correspond to equation (7). The odd \( i \) represent the wires on the top layer, and even \( i \) represent the bottom wires. The angle \( \epsilon \) in equation (7), (8), and (9) could be written as

\[ \epsilon = 2 \times \tan^{-1} \frac{W}{2L} \]  

where \( L \) is the height of each wire. Additional, the parameters \( \mu \) and \( \nu \) is expressed by

\[ \mu = \nu = \frac{(j - i - 1)}{2} \times l \]
where the $l$ is the length of each wire. Finally, the total mutual-inductance of the triangular model is summated by

\[ M_{\text{total (triangular)}} = M_{\text{vertical}} + M_{\text{horizontal (paralleled)}} + M_{\text{horizontal (non-paralleled)}} \] (18)

Figure 2.14 Geometric configuration of horizontal non-parallel wires.

Square Model

Figure 2.15 is the 3D view of the squared model which include the paralleled wires, and straight rectangular wires cases. There is no non-parallel wires in the squared model. The self-inductance and the mutual-inductance between the posts can be calculated by the equation (4b), and (5) which are already discussed in the zigzag model.
The paralleled wires have three cases which depends on the relative position between the wires. The first case include all the wires are on the Y direction and horizontal layer, as shown in Figure 2.16. The summation could be written as

\[
M_{\text{horizontal}}(Y) = \sum_{i=1}^{N-1} \sum_{j=i+1}^{N} (-1)^k M_{\text{paralleled}(i,j)}
\]  

(19)

where \(M_{\text{paralleled}(i,j)}\) correspond to the equation (6). The \(k\) is 1 if \(i + j\) is odd due to the opposite current, else \(k\) is 0. The distance \(d\) in equation (6) is represented by \(D\) and written by

\[
D = \sqrt{(k \times h)^2 + (k' \times W)^2}
\]

(20)

where \(k' = (j - i)/2\), \(h\) is the height of the post, and \(W\) is the width between each turns.

For the paralleled wires on the \(X\) direction, the total mutual-inductance is expressed as
\[ M_{\text{horizontal}}(X) = \sum_{i=1}^{N-1} \sum_{j=i+1}^{N} M_{\text{paralleled}}(i,j) \]  

(21)

The paralleled wires are on the different layers if \((i + j)\) is odd, as shown in Fig. 11.

The \(\delta\) in equation (6) is expressed by

\[
\delta' = \frac{(j - i - 1)}{2} \times \frac{W}{2}
\]

(22a)

\[
\delta'' = \frac{(j - i)}{2} \times \frac{W}{2}
\]

(22b)

Figure 2.17 Geometric configuration of X direction wires.

where \(\delta'\) is the gap between the \(i\) segment and the "a" part of the \(j\) segment, and the \(\delta''\) is the gap between the \(i\) segment and the "b" part of the \(j\) segment. The corresponding distance \(d\) in equation (6) is written respectively by

\[
D' = h
\]

(23a)

\[
D'' = \sqrt{h^2 + L^2}
\]

(23b)

Another case is \((i + j)\) is even which mean the paralleled wires on the same layers, as shown in Figure 2.17. The gap \(b\) and distance \(d\) between the two segments is written as

\[
\delta''' = \frac{(j - i - 1)}{2} \times \frac{W}{2}
\]

(24a)

\[
D''' = L
\]

(24b)
Finally, there are $2N - 1$ paralleled segments pair are missed in the last cases because they have the same number. Figure 2.18 shows these pairs which have distance $d$ in (6) equal to $L$ and the gap $\delta$ is 0. The missed mutual-inductance is written as

$$M_{\text{horizontal}}(x_{\text{miss}}) = \sum_{i=1}^{2N-1} M_{\text{paralleled}}(i,j)$$

(25)

with

$$\delta_{\text{miss}} = 0$$

(26a)

$$D_{\text{miss}} = L$$

(26b)

The total mutual-inductance in the squared model is expressed by

$$M_{\text{total(squared)}} = M_{\text{vertical}} + M_{\text{horizontal}(y)} + M_{\text{horizontal}(x)} + M_{\text{horizontal}(x_{\text{miss}})}$$

(27)

2.3. Device Fabrication

Figure 2.19 shows a general overview of the fabrication process for both the 3D inductors and resonators. The fabrication started from 3 inch high resistivity silicon wafer. The first seed layer (Au 50nm/Ti 10nm) was deposited using a DC sputtering system.
trench area is created in Si using dry etching process in an RIE system. Using standard photolithography and electroplating process, a 3 µm gold layer is deposited. BST thin film layer (250nm) was deposited using the Neocera Pioneer 180 large area pulsed laser deposition (PLD) system. The thin BST film was etched by dry etching. A second seed layer was plated on the BST thin film in order to grow the posts. A thick photoresist layer (6.4µm NR9-600PY) is deposited and etched at the posts location. The post metal (Au) was plated using the second seed layer. All the remaining NR9-600PY was cleaned by acetone. Then the NR9-600PY was coated again after the second seed layer was removed. The NR9-600PY have to be developed shortly until all top of the post was exposed. A third seed layer, which was plated on the NR9-600PY, was used for thick top metal (4.5µm Au) electroplating. The final step is to remove the exposed third seed layer.

Figure 2.20 is the cross view of the 3D inductor after fabrication. The transmission line was plated as a thick post. The thick transmission line is used to decrease the resistance in the 3D inductor that it will improve the quality factor of the 3D inductor. The BST thin film was deposited on the whole wafer except at the posts position. The additional capacitance around the corner of turns are negligible even though the dielectric constant of the BST is very large (200 ∼ 800) because the gap (6.4µm) between the top and bottom metal layer is still very high.
Figure 2.19  Summarized 3D inductors and resonators fabrication process.
2.4. Device Characterizations

2.4.1. Electromagnetic Model – Zigzag Coils and Square Coils

The electromagnetic model for zigzag coils and square coils in section 3 are built in MATLAB where the code is shown in the appendices section part A. The simulation data verified that the zigzag coils increase faster than the square coils when more turns are used in the inductor due to the larger mutual-inductance in zigzag configuration as shown in Table 2.1 and Table 2.2, especially the negative mutual-inductance of less turns square coils decrease the inductance further. The mutual-inductance difference between the two models is shown in Figure 2.21 where the mutual-inductance of the square coils become positive after 4 turns. 50 turns zigzag coils provide more than 20nH mutual-inductance than the square coils. Therefore, the zigzag model is a better choice when lots of turns structure are used in 3D inductors. The mutual-inductance of zigzag coils always increase faster than square coils because of its unique non-parallel wires structure. Figure 3.22 plot the mutual-inductance of non-parallel wires versus increased turns where the
The inductance value is close to the results of zigzag coils in Figure 3.21. The data verified the non-parallel wires structure could provide more than 90% mutual-inductance that it is a crucial factor in 3D inductor design.

<table>
<thead>
<tr>
<th>Table 2.1</th>
<th>Theoretical inductance of 3D inductors.</th>
</tr>
</thead>
<tbody>
<tr>
<td>3D inductor</td>
<td></td>
</tr>
<tr>
<td>Zigzag model</td>
<td>$L_{mutul}$ (nH)</td>
</tr>
<tr>
<td>1 turn</td>
<td>0.00637</td>
</tr>
<tr>
<td>2 turns</td>
<td>0.0343</td>
</tr>
<tr>
<td>3 turns</td>
<td>0.0826</td>
</tr>
<tr>
<td>Square model</td>
<td>$L_{mutul}$ (nH)</td>
</tr>
<tr>
<td>1 turn</td>
<td>-0.00418</td>
</tr>
<tr>
<td>2 turns</td>
<td>-0.00884</td>
</tr>
<tr>
<td>3 turns</td>
<td>-0.137</td>
</tr>
</tbody>
</table>
Table 2.2 Theoretical inductance of the inductor branch in resonators.

<table>
<thead>
<tr>
<th>Inductor branch in resonators</th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Zigzag model</td>
<td>L_{mutual} (nH)</td>
<td>wire length (μm)</td>
<td>L_{self} (nH)</td>
</tr>
<tr>
<td>1 turn</td>
<td>0.00637</td>
<td>278</td>
<td>0.257</td>
</tr>
<tr>
<td>2 turns</td>
<td>0.0343</td>
<td>306</td>
<td>0.289</td>
</tr>
<tr>
<td>3 turns</td>
<td>0.0826</td>
<td>334</td>
<td>0.321</td>
</tr>
<tr>
<td>Square model</td>
<td>L_{mutual} (nH)</td>
<td>wire length (μm)</td>
<td>L_{self} (nH)</td>
</tr>
<tr>
<td>1 turn</td>
<td>-0.00418</td>
<td>280</td>
<td>0.259</td>
</tr>
<tr>
<td>2 turns</td>
<td>-0.00884</td>
<td>320</td>
<td>0.305</td>
</tr>
<tr>
<td>3 turns</td>
<td>-0.0137</td>
<td>360</td>
<td>0.351</td>
</tr>
</tbody>
</table>

Figure 2.21 Mutual-inductance difference between zigzag coils and square coils. The mutual-inductance of zigzag coils increase faster and faster.
Figure 2.22 The inductance of non-parallel wires versus increased turns. The simulation data verified the non-parallel wires are the crucial structures in 3D inductor that it provide more than 90% mutual-inductance.

2.4.2. Simulation – 3D Inductor and Resonator

The 3D inductor and resonator models were simulated by Advanced Design System (ADS) that is a powerful 3D electromagnetic simulator. The substrate set-up in the ADS is shown in Figure 2.23. The cond and cond2 are the top and bottom conductor layers, the hole conductor represent the post in 3D inductor, and the resi conductor is the thick transmission line that is used to create the varactor in the resonator. All of the conductors are set to be made of gold. The turns for each configuration (rectangular, circle, zigzag) increase from 1 to 3. The top view and 3D layout of the 3-turns zigzag insolated inductor integrated resonators are shown in Figure 2.24 and Figure 2.25. The inductance and Q factor of the insolated inductors are calculated by the following equation [35]
\[ L = \frac{1}{2\pi f} \text{Im}(\frac{1}{Y_{21}}) \]  \hspace{1cm} (28)

\[ Q = \frac{\text{Im}(\frac{1}{Y_{11}})}{\text{Re}(\frac{1}{Y_{11}})} \]  \hspace{1cm} (29)

where \( Y_{21} \) and \( Y_{11} \) are the admittance parameters (Y-Parameters) that are converted from the scattering parameters (S-Parameters) by [36]

\[ Y_{11} = \frac{1}{Z_0} \frac{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}} \]  \hspace{1cm} (30a)

\[ Y_{12} = \frac{1}{Z_0} \frac{-2 * S_{12}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}} \]  \hspace{1cm} (30b)

\[ Y_{21} = \frac{1}{Z_0} \frac{-2 * S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}} \]  \hspace{1cm} (30c)

\[ Y_{22} = \frac{1}{Z_0} \frac{(1 + S_{11})(1 - S_{22}) + S_{12}S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}} \]  \hspace{1cm} (30d)
Figure 2.23  Substrate setting for 3D inductor simulation in ADS.

(a) Zigzag inductor model with 3 turns coil  
(b) Resonator with zigzag inductors

Figure 2.24  Top view of inductor layouts in ADS simulator. (a) Zigzag inductor model with 3 turns coil. (b) Resonator is integrated with two 3 turns zigzag inductors and a varactor.
Figure 2.25  3D view of inductor layouts in ADS simulator. (a) Zigzag inductor model with 3 turns coil. (b) Resonator is integrated with two 3 turns zigzag inductors and a varactor.

Figure 2.26  plot the inductance and Q factor of the zigzag inductors where the measurement frequency is from 0.5GHz to 20GHz. The inductance increases from 0.38nH to 0.42nH using more turns due to its increasing self-inductance and mutual-inductance. However, the Q factor decreases at the same time because of the faster increasing resistance. Figure 2.27 and Figure 2.28 exhibit the similar results of the square and circular
inductor structures respectively. The inductance difference between the three models are shown in Figure 2.29 where the zigzag coils always provide the fastest inductance increase that is match to the electromagnetic simulation in the last section. The Q factor difference is shown in Figure 2.30 where the zigzag coils have the fastest degradation because of the higher resistance of the longer wires.

Figure 2.26  Inductance and Q factor for zigzag coils configuration simulation.
Figure 2.27 Inductance and Q factor for zigzag coils configuration simulation.

Figure 2.28 Inductance and Q factor for circular coils configuration simulation.
Figure 2.29  Comparison for the inductance of different configurations. The inductance of zigzag model increases much faster than other two models.
Figure 2.30  Comparison for the Q factor of different configurations. The inductance of zigzag model have the fastest degradation due to its effective series resistance.
The insertion loss in Figure 2.31, Figure 2.32, and Figure 2.33 exhibit the characteristics of the integrated resonators with different 3D inductor models. The varactor for each model have same overlap area. The resonance point move to higher frequency and have lower amplitude using lesser turns inductors in the resonator due to the smaller inductance and higher Q factor. The resonators with zigzag inductors exhibit lowest resonance frequency because of its highest inductance and lowest Q factor.

Figure 2.31 Insertion loss of integrated resonator with shunted zigzag coils.
Figure 2.32 Insertion loss of integrated resonator with shunted square coils.

Figure 2.33 Insertion loss of integrated resonator with shunted circular coils.
2.4.3. Experiment – Resonator

The microphotograph for the fabricated integrated resonators based on 3D inductors and varactors on a high resistivity Si substrate are shown in Figure 2.34. As it can be clearly seen the filter structures with T-line (no coils), 1 squared turn, and 3 squared turns.

The S-parameters of the resonators are measured using a vector network analyzer and is illustrated further using an equivalent schematic circuit based on a basic T model in AWR simulator, as shown in Figure 2.35. The transmission line segments are present automatically by the CPW model which is existing in the AWR simulator. The varactor value is modeled by a tunable capacitor C and the resistance in the BST materials is modeled by a huge resistor \( R_C \). The inductance and series resistance of the two shunt coils are performed by a single inductor L and resistor \( R_L \) respectively. The Q factor of the capacitor and inductor are calculated from the S parameters by equations

\[
Q_L = \frac{\omega_0 \times L}{R_L} \quad (28a)
\]

\[
Q_C = \omega_0 \times C \times R_C \quad (28b)
\]

where \( \omega_0 = 2\pi f \) is the angular frequency.

The experimental insertion loss, at frequency range from 0GHz to 20GHz, are presented in Figure 2.36, Figure 2.37, and Figure 2.38 for the resonators with different inductor configurations. The resonance frequency of resonator with T-line segments is 9.96GHz and the 3dB cut-off frequency is \( \sim 4.6GHz \). The resonance frequency of resonator with 1 turn coil decrease to 9.41GHz due to the increasing inductance from the coil, then the cut-off frequency decreased to 4.4GHz. The resonance frequency continually decreases to 8.6GHz for resonator with 3 turns coils, and the cut-off frequency decreased to 4.27GHz.
The schematic model parameters, as shown in Table 2.3, are fitted to the measured insertion loss.

Table 2.3 The schematic model parameters used to fit the experiment results.

<table>
<thead>
<tr>
<th>Device</th>
<th>C ($pF$)</th>
<th>$R_C$ ($\Omega$)</th>
<th>$Q_C$</th>
<th>$L$ ($nH$)</th>
<th>$R_L$ ($\Omega$)</th>
<th>$Q_L$</th>
</tr>
</thead>
<tbody>
<tr>
<td>T-line</td>
<td>0.9</td>
<td>500</td>
<td>28</td>
<td>0.28</td>
<td>4.383</td>
<td>4</td>
</tr>
<tr>
<td>1 turn</td>
<td>0.9</td>
<td>500</td>
<td>26.9</td>
<td>0.318</td>
<td>4.282</td>
<td>4.43</td>
</tr>
<tr>
<td>3 turns</td>
<td>0.9</td>
<td>500</td>
<td>24.5</td>
<td>0.368</td>
<td>5.279</td>
<td>3.85</td>
</tr>
</tbody>
</table>

The equivalent resonators are compared in Figure 2.39 where the resonance points move to lower frequency and lower amplitude using more turns inductors because of the increasing inductance that is already learned from both the electromagnetic and simulation results. Unfortunately, the resonance point didn't match very well with the simulation results. The reason is undesired BST thin film growth due to our beam scanning system out-of-alignment. The fabricated BST have much smaller dielectric constant than the usual BST [6]. Therefore, the resonance point moved to higher frequency.
Figure 2.34  Microphotograph view of fabricated devices. (a) Resonator with T-line. (b) Resonator with two shunted 1 turn squared coils. (c) Resonator with two shunted 3 turns squared coils. The bottom pictures show the details at the 3D inductor coils.

Figure 2.35  The equivalent circuit model of the resonator.
Figure 2.36  The experimental resonator versus the equivalent resonator with shunted T-line branches.

Figure 2.37  The experimental resonator versus the equivalent resonator with shunted 3 turns squared coils.
Figure 2.38  The experimental resonator versus the equivalent resonator with shunted 1 turn squared coils.

Figure 2.39  The equivalent resonator using different shunted branches. More turns will make resonance points move to lower frequency and amplitude.
CHAPTER III
TUNABLE INDUCTORS USING VANADIUM DIOXIDE AS CONTROL MATERIAL

This chapter starts with a review of the VO$_2$ thin film and an existing tunable inductor [15] in section 3.1. In section 3.2, the configuration and working principal of each design are studied. The method of a complex fabrication for the VO$_2$ based actuator and a simplified fabrication process for the tunable inductor using VO$_2$ as switches are introduced in section 3.3. In section 3.4, the simulated and experimental results of each design are analyzed in terms of the equivalent inductance and equivalent circuit.

3.1. VO$_2$ and Existing Tunable Inductor Review

3.1.1. Vanadium Dioxide (VO$_2$)

The study of the Vanadium dioxide started in 1959 when Morin observed the metal-to-insulator transition (MIT) characteristics in VO$_2$ [37]. VO$_2$ is a kind of phase change material that it transitions from insulator to conductor in electrical property when temperature rises past the transition temperature due to its crystal structure changes from monoclinic to tetragonal above the transition temperature [38]–[44]. VO$_2$ is widely used in reconfiguration components and circuits in the last decade because of its transition temperature are much lower than other vanadium oxides materials such as
V₆O₁₃, V₂O₅, and V₂O₅. Abrupt mechanical displacement of VO₂ beams is generated at transition temperature too because of the MIT created surface stress change on the VO₂ beam [45], [46].

Rafmag [24] tested a VO₂-based actuator performance recently, as shown in Figure 3.1. The maximum deflections of a 300 µm cantilevers were raised from 73µm at room temperature to 142µm at 470 °C. The 200 µm cantilevers only could raise from 42µm to 70µm. However, the sudden movement (abrupt displacement) of VO₂ cantilevers reduced the tuning range of the actuator. The cantilevers could only stop at the original position and the highest position. It was very hard to stop the VO₂ cantilevers when they were moving. Furthermore, the displacement is still one direction (upward). Continuous inductance tuning is difficult to achieve in this design.

Figure 3.1 Cross-section view of a simple VO₂ actuator with varied temperature in [24]. (a) 200µm VO₂ cantilevers. (b) 300µm VO₂ cantilevers.
3.1.2. Existing Tunable Inductor

In 2009, Kim [15] fabricated a tunable inductor with large displacement obtained using an electro-thermal actuator. The inductance was tuned by a magnetically coupled short-circuit loop that its position was controlled by an electro-thermal actuator. Control of the distance between the spiral coil and the circuit loop could tune the mutual-inductance. However, the cantilevers only moved downward when Ti cantilevers were heated up that it is still one direction displacement. The 3D view of the tunable inductor is shown in Figure 3.2.

![Configuration of existing tunable inductor with bimorph actuator in [15].](image_url)
3.2. Methods

3.2.1. Tunable Inductor using VO$_2$ as Actuator

The integrated actuator with both metal and VO$_2$ cantilevers (mechanical displacement) are applied in the first tunable inductor design. The novel actuator is expected to provide both up and down movement, gradually, resulting in continuous tuning of the inductance. The VO$_2$ cantilevers can raise the actuator to the high position when they are heated to pass the transition temperature. The heated metal cantilevers can pull the actuator down gradually. After releasing the heating on VO$_2$ cantilevers, the metal actuator move down continually to the low position. The tunable inductors are combined with a magnetic core that it is above a conventional planar coil inductor. The magnetic core interaction with the coil is controlled by the integrated actuator. Inductance variation is achieved by changing the relative position of magnetic material which changes the permeability of inductor coils. The cross-sectional view of the whole tunable inductor is shown in Figure 3.3.

![Cross view of the novel tunable inductors.](image.png)
One 200 µm VO$_2$ cantilever is coated in between two or four 200 µm Au cantilevers, as shown in Figure 3.4. The heater of VO$_2$ beam is designed as a plane coil which is only 5 µm wide. Another 100 µm VO$_2$ cantilever is also be used in the actuator to compare the tunability of different length of VO$_2$ beam. The VO$_2$ cantilever is 40 µm wide, and the Au cantilevers are 5 µm wide, same as the VO$_2$ heaters. The narrow coil Au wires can increase the resistance of the cantilevers in order to concentrate most of power at the coil part. Therefore, the coil structure can be used for both cantilevers and heaters. A 350 µm×350 µm Nickel pad, which has a permeability of close to 100 [47], is coated at the end of the SiO$_2$ anchor which is used to hold the whole actuator. The conventional coil inductor are designed with Ground/Signal/Ground (GSG) coplanar waveguide (CPW) transmission line feed structures on either side. The center signal line is 50 µm wide and the separation gap between the signal line and the ground is also 50µm. The coil inductor use 10 µm wires with 50 µm, 35 µm, and 20 µm space, as shown in Figure 3.5.

![Figure 3.4 Top view of the VO$_2$ based actuator.](image-url)
3.2.2. Tunable Inductor using \( \text{VO}_2 \) as Switches

For the second tunable inductor design, two configurations are presented using the \( \text{VO}_2 \) thin film that can control part of the inductor coil due to the \( \text{VO}_2 \) metal-to-insulator transition characteristics. A single \( \text{VO}_2 \) thin film is used in this design to replace the traditional switches in other tunable inductors which require complicated fabrication, high manufacturing costs, larger size, and have short lifetime [5]–[7]. The \( \text{VO}_2 \) thin film deposition is done using a PLD process which is a much simplified fabrication. A functional and reliable device like this would have good benefits such as size reduction, long lifetime, and decreased cost to manufacture.

A short section of \( \text{VO}_2 \) are used in the first configuration as shown in Figure 3.6(a) where the \( \text{VO}_2 \) acting as a switch at different temperature. For the second configuration in Figure 3.6(b), the Au inductor are considered to be connected in parallel with a \( \text{VO}_2 \) acting as an
effective inductor when device is heated higher than 70°C. Both the Au and VO$_2$ spiral coil are 10µm wide and have 50µm gap space between the spiral wires. The VO$_2$ bar is 20µm wide to reduce the resistance in the transmission line and 170µm long to connect to the bottom Au inductor spiral.

At low temperatures e.g. room temperature, only the top single spiral works since the VO$_2$ is at insulating state with the high resistance blocking signals from passing through the bottom spiral. The dual spiral structure begins to work at high temperatures e.g. 80 °C and above when the VO$_2$ enters conducting state with low resistance to allow signals to pass through the bottom spiral. The inductance of the device should be ideally halved at high temperature operation.

Figure 3.6   Microphotograph of the tunable inductors with (a) VO$_2$ bar and (b) VO$_2$ spiral coil. Light yellow wires are the bottom gold layer. Brown wires are the VO$_2$ layer.
3.2.3. Multi-tap Inductors using VO$_2$ as Switches

Two configurations of multi-tap inductors using VO$_2$ thin films are designed. In the first configuration, four output ports are used as shown in Figure 3.7(a) where the ports are connected with CPW transmission lines using zigzag turns and controlled using VO$_2$ bars. These short VO$_2$ sections act as switches controllable using temperature. The inductor is designed using 30μm/70μm/30μm CPW transmission line due to its 50Ω input impedance. However, the impedance mismatch are presented at the zigzag turns because of the varied width and gap. In order to reduce the impedance mismatching in the transmission line, step width junctions are used in the second configuration as shown in Figure 3.7(b) where the width and gap variation are disappeared. Simultaneously, the discontinuity effects of the step width junctions also create fringing capacitances that are not negligible at RF and microwave frequency. The device is turned off when the VO$_2$ is insulating at low temperature. After the VO$_2$ thin film transitions to conducting state above the transition temperature, the signal pass through the device and output from the port which is connected with the output probe. The inductance of the device is determined by the length of the transmission line.
Figure 3.7  Microphotograph of the multi-tap inductors with (a) zigzag turns in the first model and (b) step width junctions in the second model. The prot-5 is connected with device using Au.

3.3. Device Fabrication

3.3.1. VO$_2$ based Actuator

The VO$_2$ actuator is a crucial unit in this tunable inductor design. Figure 3.8 shows a general overview of the fabrication process. The starting substrate consisted of a 500 $\mu$m thick silicon wafer where the planar coil inductor is already existing on it. A 1 $\mu$m polysilicon film and 500 nm SiO$_2$ film anchor are deposited using sputter deposition process and patterned using a dry etch process. A 150 nm VO$_2$ thin film is deposited on the SiO$_2$ anchor by PLD at 500°C and patterned using a dry etch process. A 300nm Au cantilever and heater layer are deposited on the SiO$_2$ anchor using e-beam evaporation. The Au layer has to be coated on top of the VO$_2$ cantilevers to avoid the high temperature VO$_2$
deposition process which may soften and melt the Au cantilevers. Finally, a 500 nm Nickel layer is deposited on the SiO$_2$ anchor, and the bare poly-silicon film is etched using a dry etch process.

![Fabrication process flow of the actuator for.](image)

**3.3.2. Tunable Inductor using VO$_2$ as Switches**

Figure 3.9 shows a general overview of fabrication process for the second tunable inductor. The devices fabrication is started on 3 inch double side polished c-cut Sapphire wafer. A 320 nm metal layer (Ti 20 nm/Au 200 nm/Pt 100 nm) is deposited on the substrate by e-beam evaporation and lift-off process. A 500 nm SiO$_2$ thin film is deposited by PECVD and patterned using a dry etch process. The via structure is filled by another metallization (Ti 20nm/Au 300nm/Pt 200nm). A 150nm thick VO$_2$ thin film is deposited by
Pulsed Laser Deposition at 500°C [48] and patterned by dry etch using SF₆. The final step is the top metal layer (Ti 20nm/Au 300nm) deposition and lift-off process.

The finalized VO₂ thin film experimental measurements are shown in Figure 3.10 where it include the measured XRD of the wafer and the SEM image of the VO₂ in (a), and the measured resistivity of the VO₂ thin film in (b). The VO₂ film is formed along (0, 2, 0) on a SiO₂ surface compared to (0, 1, 1) on a sapphire surface in [48]. The VO₂ quality is verified by a test structure as shown in Figure 3.10(b) before the tunable inductors are measured. The test structure is a traditional CPW transmission line with a 10 μm × 50 μm gap which is filled with VO₂ on a 500 nm SiO₂ thin film substrate. Resistivity measurements in Figure 3.10(b) show a hysteresis loop between the heating up and cooling down transitions. However, the tuning resistivity ratio still remains high. For the test structure, the resistivity ratio is ~ 10000:1 (Resistance from 300KΩ to 25Ω) from room temperature to 100°C where a rapid switching is observed around 70°C. Figure 3.11 presents the SEM and the measured resistivity of a similar test structure which is filled VO₂ on a sapphire substrate. A narrow hysteresis loop is observed with no significant change between the heating up and cooling down transitions. The resistivity ratio is ~ 15000:1 (Resistance from 130KΩ to 9Ω) that it close to a pure conductor after the temperature rise above 70°C. The insertion loss curves in Figure 3.12 presents nice working MIT behavior when the VO₂ is heated from 25°C to 100°C. Most of the signal is blocked at room temperature, but more than 80% of the signal is able to pass through the device at 100°C.
Figure 3.9 Fabrication process for tunable inductor using VO$_2$ as switches.
Figure 3.10  VO$_2$ characteristics on SiO$_2$ surface. (a) XRD and SEM of the VO$_2$ thin film. (b) Resistance of VO$_2$ by using variation temperature (logarithm scale).
Figure 3.11  VO₂ characteristics on Sapphire surface. (a) SEM of the VO₂ thin film. (b) Resistance of VO₂ by using variation temperature (logarithm scale).
3.4. Devices Characterizations

3.4.1 Tunable Inductor using VO₂ as Actuator

The analysis and modeling of the resulting structures are performed from 0.3 GHz to 20 GHz using Advanced Design System (ADS) simulator. The substrate set-up in ADS is shown in Figure 3.13. A 1 μm Nickel magnetic layer is covered on a coil inductor with a relative permeability of 100. The gap between the coil inductor and Nickel layer are changed from 5 μm to ∞ (no Nickel layer) to simulate the Nickel is moved by a VO₂-based actuator. The VO₂, Au cantilevers, and Nickel pad are plated on a SiO₂ anchor. The tunability is calculated by the following equations [35].
Equivalent inductance \( L = \frac{\text{imag}(-\frac{1}{Y_{21}})}{2\pi f} \) \hfill (1)

Tunability \( = \frac{L_{\text{high}} - L_{\text{low}}}{L_{\text{high}}} \times 100\% \) \hfill (2)

where \( L_{\text{high}} \) is the inductance when the Nickel is far away from the inductor, and \( L_{\text{low}} \) is the inductance when the Nickel is close to the inductor.

The inductance decreases from 2.43nH to 1.91nH (27% inductance tunability) at 5 GHz as we increase the Nickel layer height from 5 \( \mu m \) to \( \infty \) on a coils with 35 \( \mu m \) wire space due to the effective permeability of the coil inductor become larger by the far Nickel layer, as shown in Figure 3.14. The inductance tunability is shown in Figure 3.15 where the simulated inductance variation at 5GHz is approximately 27%. A \( \sim 22\% \) decreased inductance is exhibited when the air gap change from 5 \( \mu m \) to 50 \( \mu m \). Furthermore, the tunability does not change by the varied wire space which mean the inductance variation range will become larger using larger inductance. The Q factor, as shown in Figure 3.16 and Figure 3.17, also become larger when the Nickel layer move close to the coil inductors because of the increased inductance and constant resistance of the coils. The highest Q factor move to lower frequency. Wider space coil can reduce the Q factor due to its larger inductance.
Figure 3.13  Substrate setup for tunable inductor using VO$_2$-based integrated actuator.

Figure 3.14  Simulated inductance for varied air gap between the Nickel and the coils.
Figure 3.15 Inductor tunability versus air gap.

Figure 3.16 Simulated Q factor for varied air gap between the Nickel and coils.
3.4.2 Tunable Inductor using VO$_2$ as Switches

Two configurations of tunable inductor using VO$_2$ as switches are measured on an on-wafer probe station with a non-ground heater that its temperature is controlled by a thermoelectric temperature controller from 20°C to 100°C. The S parameters of the devices are collected using a network analyzer with 150μm GSG probes from 0.3 GHz to 2 GHz. The collected data are converted to Y-parameters using the AWR simulator or using the equation (30), and then the inductance tunability are written as

$$Tunability = \frac{L_{low} - L_{high}}{L_{low}} \times 100\%$$

(3)
Where $L_{high}$ is the inductance at high temperature, and $L_{low}$ is the inductance at low temperature.

Figure 3.18 plots the inductance and tunability versus varied temperature for the tunable inductor with a VO$_2$ short bar where a significant change occur between 60°C and 80°C due to VO$_2$ phase change beyond the transition temperature. The inductance at 1GHz decrease from 1nH to 0.68nH that the tunability is $\sim 32\%$ which is lower than 50% of the idea case for two same inductor branches because the VO$_2$ have higher resistivity than Au over the transition temperature. About 23% inductance change occur between 60°C and 80°C. Figure 3.19 plots the inductance and tunability versus varied temperature for the tunable inductor with a VO$_2$ coil acts as an inductor where the inductance variation at each temperature stage are same. The tunability is $\sim 26\%$ (1nH to 0.76nH) is smaller than the first configuration due to the longer VO$_2$ wire generate larger resistance in the bottom inductor branch and inconsistent VO$_2$ quality. The abrupt inductance change still exists around the transition temperature, but it is not as significant as in the first design.

Both two configurations have undesired Q factor, which is smaller than 1, because the high resistance from the thin combined Au metal layer (320nm) and long VO$_2$ wire (>150μm). Most power is lost on the Au and VO$_2$. Thick metal (>2μm), thick VO$_2$ thin film (>300nm), and short VO$_2$ wire (<20μm) will be used in the future design to improve the Q factor. Simultaneously, better quality VO$_2$ also is necessary for the future design.
Figure 3.18 The (a) inductance variation and (b) tunability of inductor with VO$_2$ bar versus temperature. A rapid change is happened at 60°C – 80°C temperature stage.
Figure 3.19 The (a) inductance variation and (b) tunability of inductor with VO₂ spiral coil versus temperature. Gradual and slow inductance variation is exhibited.
An equivalent circuit model of the proposed tunable inductor, as shown in Figure 3.20, is illustrated based on a basic \( \pi \) model where the inductance is controlled by a varied VO\(_2\) based resistor \( (R_{VO_2}) \). Only the first VO\(_2\) short bar configuration is modeled in this section due to its better tunability ratio and conforming VO\(_2\) quality. The measured insertion loss and inductance is fitted using the parameters in Table 3.1. The oxide capacitance between the coils and the substrate is modeled by \( C_{ox} \) and \( C_{sub} \), and the capacitance between the spiral wires are modeled by \( C_s \). The inductance of the gold coils \( (L_{spiral}) \) is constant in the schematic model that it is equal to the inductance of the tunable inductor at room temperature in Figure 3.18(a). The resistance of the gold spiral coil \( (R_{spiral}) \) increase 14\( \Omega \) at 100\( ^\circ \)C because of the temperature coefficient of the gold wires. The resistance of VO\(_2\) bar \( (R_{VO_2}) \) decrease from hundreds of kilo-ohms to 180\( \Omega \) after the devices are heated from room temperature to 100\( ^\circ \)C that the resistance value is conform to the measured resistivity results of the test structure. At low temperature, the effective inductance is equal to \( L_{spiral} \) due to the signal is blocked by the huge \( R_{VO_2} \). The effective inductance decrease at low temperature when signal start to pass thorough the bottom inductor branch because of the smaller \( R_{VO_2} \). The insertion loss and inductance in Figure 3.21 show a strong matching between the circuit model and the experimental results.
Figure 3.20 Equivalent circuit model of inductor with VO₂ bar structure.

Table 3.1 The schematic model parameters to fit the experimental results.

<table>
<thead>
<tr>
<th></th>
<th>( R_{VO₂} )</th>
<th>( R_{spiral} )</th>
<th>( L_{spiral} )</th>
<th>( C_{ox} )</th>
<th>( C_{sub} )</th>
<th>( R_{sub} )</th>
<th>( C_s )</th>
</tr>
</thead>
<tbody>
<tr>
<td>25°C</td>
<td>&gt;300KΩ</td>
<td>39Ω</td>
<td>1nH</td>
<td>0.8pF</td>
<td>0.8pF</td>
<td>&gt;10KΩ</td>
<td>10fF</td>
</tr>
<tr>
<td>100°C</td>
<td>180Ω</td>
<td>53Ω</td>
<td>1nH</td>
<td>0.8pF</td>
<td>0.8pF</td>
<td>&gt;10KΩ</td>
<td>10fF</td>
</tr>
</tbody>
</table>
Figure 3.21  (a) Inductance variation and (b) tunability of inductor with VO$_2$ bar structure versus temperature. Gradual and slow inductance variation is exhibit.
3.4.3 Multi-tap Inductor using VO₂ as Switches

Two configurations of multi-tap inductors are tested on a probe station, and the S parameters are collected using network analyzer and two 150µm GSG probes. A thermoelectric temperature controller is used to change the devices temperature from 20°C to 80°C.

Figure 3.22 and Figure 3.23 plot the measured insertion loss $S_{21}$ where the dash lines are compared devices that only Au wires are used at each port. When the output probe is moved and connected with different output ports, a clear insertion loss variation is created that the resonance move to higher frequency. The strong matching between the VO₂ based devices and the Au based devices reveal that its quality is better than the old thin film in the previous research [48]. The insertion loss versus temperature in Figure 3.22(b) and Figure 3.23(b) present that more than 90% signal are blocked by the VO₂ in its insulating state at room temperature and at least 80% signal pass through the device and output from port-1 by the VO₂ conducting state at 80°C. Figure 3.24 plot the inductance of the device at 80°C using the equation

$$\text{Equivalent inductance } L = \frac{\text{imag}(-\frac{1}{Y_{21}})}{2\pi f} \tag{4}$$

where the transfer admittance $Y_{21}$ is converted from S parameters using AWR simulator.

The inductance of the first model increase from 0.57nH to 1.82nH gradually at 2 GHz when the output probe is moved from port-1 to port-5. The inductance of the second model decrease from 1.24nH to 1nH when the probe move from port-1 to port-2, then the inductance increase from 1nH to 1.82nH when the probe move from port-2 to port-5. The
second design exhibit stronger capacitive character and worse inductive variation at high frequency due to the step width junctions providing more fringing capacitances and discontinuity effects in the devices.

Figure 3.22 The insertion loss of multi-tap inductor with zigzag turns for each port at 80°C (a) and the port-1 versus varied temperature (b). The VO₂ exhibit good quality at high temperature that its conductivity is close to Au.
Figure 3.23  The insertion loss of multi-tap inductor with step width junction for each port at 80°C (a) and the port-1 versus varied temperature (b). VO₂ exhibit good quality that its conductivity is close to Au.
Figure 3.24  Equivalent inductance of the (a) zigzag turns model and (b) step width junction model. Stronger inductive character and stable inductance variation are present in the first design.
The high temperature multi-tap inductor model is illustrated further by using an equivalent circuit based on CPW circuit elements as shown in Figure 3.25 where the inductance is controlled by the moving output port which decide the effective length of the transmission line in the device. The coplanar open CPW elements are used in the device to simulate the unused open ports (port-2 to port-5). The transmission lines between the ports are modeled using four lengths of CPW elements ($CPW_{input} = 650 \mu m$, $CPW_{medium} = 690 \mu m$, $CPW_{VO2} = 200 \mu m$, and $CPW_{Au} = 890 \mu m$), and the 20 \mu m long VO2 thin film is modeled using a normal resistor ($R_{VO2} = 5\Omega$). The insertion loss in Figure 3.26 show a strong matching between the circuit model and the experimental results.

Figure 3.25   Equivalent circuit model of the multi-tap inductor with zigzag turns.
Figure 3.26 The insertion loss of the equivalent circuit model and the experimental results. A strong matching is exhibited.
4.1. Conclusions

A 3D inductor with three different configurations were designed, fabricated and measured. A resonator combining the 3D inductors and BST thin film varactor is also verified for the first time. A complex fabrication process is developed to create an effective approach to utilize the 3D inductors in electrical applications such as resonators and filters. The electromagnetic models are verified using MATLAB. The zigzag configuration always provide the biggest mutual-inductance due to its unique non-parallel structures, and its mutual-inductance also increase much faster than other configurations. The inductance of except self-inductance of zigzag coils can achieve to 20nH when more than 50 zigzag turns are used in the inductors. The other configurations can only get 2nH added inductance for 50 turns inductors. The simulation of the 3D inductors show that more turns results in higher inductance and lower quality factor because the resistance of the added turns increase faster than the inductance. Therefore, the zigzag coils have the largest inductance and the lowest quality factor. The resonance frequency of the fabricated resonator correlated to the different number of turns. The equivalent schematic circuit was built with a simple capacitor and inductor. Each inductor is about
0.6nH which is higher than the simulation results. The measurement results and simulation results didn't match well because of the non-optimal BST thin film.

Two types of tunable inductors using vanadium dioxide as control materials are presented in this dissertation. The new actuator in the first design improve the performance of the existing actuator. The second tunable inductor design leads a much simplified fabrication process that would have good benefits such as overall system versatility, size reduction and decreased manufacturing. The VO$_2$ metal-to-insulator transition (MIT) and displacement characteristics are studied and used in these two types of tunable inductors respectively with $\sim 10000:1$ resistance ratio between room temperature and 80°C. The simulation data of tunable inductor using VO$_2$-based actuator exhibit at least 27% gradually inductance variation in the first design. These devices are controlled by varied permeability when a magnetic material such as nickel is moved over the inductors using the integrated actuator. Therefore, the Q factor does not degrade as the inductance increases due to the constant resistance of the inductor coils. Compared with the existing actuator in tunable inductor research, we present a two-way (bimorph) displacement actuator based on VO$_2$ and metal cantilevers to provide larger displacement.

In the second design, the tunable inductors using VO$_2$ as switches exhibit 32% inductance variation using a short VO$_2$ bar and 26% variation using a full VO$_2$ spiral. A circuit model based on the bar configuration is constructed due to its stronger VO$_2$ tuning capability and we obtain a good fit over the entire measured frequency range. The VO$_2$ exhibit good conductivity at high temperature but it is still well behind the conductivity
of Au. We present a tunable inductor with simplified fabrication, small device size, and low cost to manufacture for potential applications in portable and adaptive communication systems.

Two configurations of VO₂ based multi-tap inductors are designed, fabricated and tested in the last section. The strong matching between the devices using VO₂ thin films and the devices using Au show that we are able to produce better quality VO₂ than before [48]. The zigzag turns design exhibits stronger inductor character and better inductance variation than the step width junctions design in the RF and microwave frequency due to the discontinuity effects of the step width junctions such as fringing capacitors. A circuit model based on the zigzag turns design is constructed and results are illustrated to verify the effectiveness along with experimental data.

4.2. Future Work

In order to get better quality factor on 3D inductors, more turns should be used in zigzag models due to its biggest mutual-inductance. We believe both the inductance and quality factor will increase when more than 10 turns are used. The normal BST thin film will be used to correct the measurement results. The working varactor can make the resonator to be tunable. The DC bias will also be added on the resonator to achieve better tunability.

The tunable inductor based on the VO₂-based actuator will be fabricated and tested using a network analyzer on an on-wafer probe station with the temperature of the chuck controlled by a thermoelectric temperature controller. The Au cantilevers are controlled using DC biasing through the probes. In order to get better tunability and Q factor in the
future, a shorter VO$_2$ bar and thick metal will be used in the second design to reduce the resistance of devices. Furthermore, a shorter VO$_2$ bar also provide better phase change property and stable quality.

4.3. Publications


REFERENCES


[47] [Online]: http://hyperphysics.phy-astr.gsu.edu/hbase/tables/magprop.html#c2.

% Inductance for Zigzag Model.

clear all;

W = 21*10^-6; % width between each turns.
L = 22*10^-6; % length of vertical wires.
l = 25*10^-6;
er=2*10^-7; % permeability.
h=10*10^-6; % hight of the via.
epsilon=2*atan(W/2/L); % angle between the non-parallel wires.
phi=pi/2-epsilon/2; % angle between the wires and the horizontal axis.
Tline=278*10^-6; %Total length of all wires.
B=5*10^-6; % width of the wires.
C=4*10^-6; % thickness of the wires.
N=1; % # of turns.

% mutual-inductance between vertical wires.

M=2*N;
for i=1:M-1
    for j=i+1:M;
        kp=(j-i)/2;
        if mod((i+j),2)==0;
            k=0;
        else
            k=1;
        end
    end
end

APPENDIX A
MATLAB CODE FOR 3D INDUCTOR SIMULATION

Inductance for Zigzag Model

% Inductance for Zigzag Model.

% parameters.
clear all;

W = 21*10^-6; % width between each turns.
L = 22*10^-6; % length of vertical wires.
l = 25*10^-6;
er=2*10^-7; % permeability.
h=10*10^-6; % hight of the via.
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Tline=278*10^-6; %Total length of all wires.
B=5*10^-6; % width of the wires.
C=4*10^-6; % thickness of the wires.
N=1; % # of turns.

% mutual-inductance between vertical wires.

M=2*N;
for i=1:M-1
    for j=i+1:M;
        kp=(j-i)/2;
        if mod((i+j),2)==0;
            k=0;
        else
            k=1;
        end
    end
end
\[ d = \sqrt{(k p w)^2 + (k L)^2}; \]

if \( h/d \geq 10; \)
\[ M_{\text{vertical}}(i, j) = e r * h^* \left( \log(2*h/d) - 1 + d/h - 0.25*(h^2/d^2) \right); \]
else
if \( h/d \leq 0.1; \)
\[ M_{\text{vertical}}(i, j) = e r * h^* \left( 0.5*h/d \right) * \left( 1 - \frac{1}{12}*(h^2/d^2) + \frac{1}{40}*(h^4/d^4) \right); \]
else
\[ M_{\text{vertical}}(i, j) = e r * h^* \left( \log(h/d + \sqrt{1+h^2/d^2}) - \sqrt{1+h^2/d^2} + d/h \right); \]
end
end
\[ M_{\text{vertical}}(i, j) = (-1)^k * M_{\text{vertical}}(i, j); \]
end
end

\[ \text{/* mutual-inductance between parallel wires.} \]
\[ M = 2^N; \]
\[ \text{for } i=1:M-1 \]
\[ \text{for } j=i+1:M; \]
\[ k p = (j-i)/2; \]
if \( \text{mod}(i+j), 2 \) == 0;
\[ k p p = 1; \]
else
\[ k p p = 0; \]
end
\[ d_p = k p * w * \sin(\phi); \]
\[ \text{delta} = -(l-k p * w * \cos(\phi)); \]
\[ \alpha lpha = l+\text{delta}; \]
\[ \beta eta = \text{delta}; \]
\[ \gamma mma = \text{delta}; \]
\[ M_{\text{parallel}}(i, j) = e r / 2^* \left( \alpha lpha * \text{asinh}(\alpha lpha / d_p) - \beta eta * \text{asinh}(\beta eta / d_p) \right) - \gamma mma * \text{asinh}(\gamma mma / d_p) + \delta delta * \text{asinh}(\delta delta / d_p) \right) - \sqrt{\alpha lpha^2 + d_p^2} + \sqrt{\beta eta^2 + d_p^2} + \sqrt{\gamma mma^2 + d_p^2} - \sqrt{\delta delta^2 + d_p^2}); \]
\[ M_{\text{parallel}}(i, j) = k p p * M_{\text{parallel}}(i, j); \]
end
end

\[ \text{/* mutual-inductance between unparralle wires.} \]
\[ M = 2^N; \]
\[ \text{for } i=1:M-1 \]
\[ \text{for } j=i+1:M; \]
\[ \mu = (((j-i-1)/2)^*1; \]
\[ \nu = (((j-i-1)/2)^*1; \]
if \( \text{mod}(i+j), 2 \) == 0;
\[ k = 0; \]
else
\[ k = 1; \]
end
R1=h^2+(mu+l)^2+(nu+l)^2-2*(mu+l)*(nu+l)*cos(epsilon);
R2=h^2+(mu+l)^2+(nu+l)^2-2*(mu+l)*nu*cos(epsilon);
R3=h^2+(mu)^2+(nu)^2-2*mu*nu*cos(epsilon);
R4=h^2+(mu)^2+(nu+l)^2-2*mu*(nu+l)*cos(epsilon);

Omega=atan(((h^2)*cos(epsilon)+(mu+l)*(nu+l)*(sin(epsilon))^2)/(h*R1*sin(epsilon)))-
atan(((h^2)*cos(epsilon)+(mu+l)*nu*(sin(epsilon))^2)/(h*R2*sin(epsilon)))+atan(((h^2)
*cos(epsilon)+mu*nu*(sin(epsilon))^2)/(h*R3*sin(epsilon)))-
atan(((h^2)*cos(epsilon)+mu*(nu+l)*(sin(epsilon))^2)/(h*R4*sin(epsilon)));

M_non-
parallel(i,j)=er*cos(epsilon)*(((mu+l)*atan(l/(R1+R2))+(nu+l)*atan(l/(R1+R4))-
mu*atan(l/(R3+R4))-nu*atan(l/(R2+R3)))-Omega*h/sin(epsilon));

% total mutual inductance.
L_mutual=(sum(M_vertical(:))+sum(M_parallel(:))+sum(M_non-parallel(:)));

% self inductance.
L_self=er*Tline*(log(2*Tline/(B+C))+0.5-0.00181);

% total inductance
L_total=(L_self+L_mutual);
h=10*10^-6; \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ % hight of the via.

Tline=245*10^-6; \ \ \ \ \ \ \ \ %Total length of all wires.

B=5*10^-6; \ \ \ \ \ \ \ \ \ \ \ \ \ \ % width of the wires.
C=4*10^-6; \ \ \ \ \ \ \ \ \ % thickness of the wires.

N=1; \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ \ % # of turns

% mutual-inductance between vertical wires.

M=2*N;
for i=1:M-1
    for j=i+1:M;
        kp=(j-i)/2;
        if mod((i+j),2)==0;
            k=0;
        else
            k=1;
        end
        d=sqrt((kp*W)^2+(k*L)^2);
        if h/d >=10;
            M_vertical(i,j)=er*h*(log(2*h/d)-1+d/h-0.25*(h^2/d^2));
        else
            if h/d <= 0.1;
                M_vertical(i,j)=er*h*(0.5*h/d)*(1-(1/12)*(h^2/d^2)+(1/40)*(h^4/d^4));
            else
                M_vertical(i,j)=er*h*(log(h/d+sqrt(1+h^2/d^2))-
sqrt(1+h^2/d^2)+d/h);
            end
        end

    end

M_vertical(i,j)=(-1)^k*M_vertical(i,j);
end

% mutual-inductance between horizontal L wires.

M=2*N;
for i=1:M-1
    for j=i+1:M;
        kp=(j-i)/2;
        if mod((i+j),2)==0;
            k=0;
        else
            k=1;
        end
        D=sqrt((k*W)^2+(kp*h)^2);
        if L/D >=10;
            M_L(i,j)=er*L*(log(2*L/d)-1+d/L-0.25*(L^2/d^2));
        else
            if L/D <= 0.1;
\[ M_L(i,j) = \begin{cases} \varepsilon_r L \left(0.5L/d\right) \times (1 - (1/12)(L^2/d^2) + (1/40)(L^4/d^4)) \\ \varepsilon_r L \log(L/d) + \sqrt{(1 + L^2/d^2)} + d/L \end{cases} \]

\[
M_L(i,j) = (-1)^k M_L(i,j)
\]

\% mutual-inductance between horizontal w wires (same layers).

\[ M = 2 * N; \]

\[ \text{for } i = 1: M-1; \]
\[ \text{for } j = i+1: M; \]
\[ \text{if } \text{mod}((i+j),2) == 0; \]
\[ D_{ppp} = L; \]
\[ \delta_{ppp} = (j-i-1)/2 * (W/2); \]
\[ \alpha_{ppp} = l + \delta_{ppp}; \]
\[ \beta_{ppp} = l + \delta_{ppp}; \]
\[ \gamma_{ppp} = l + \delta_{ppp}; \]
\[ M_{W_s_{ppp}}(i,j) = \frac{\varepsilon_r}{2} (\alpha_{ppp} \text{asinh}(\alpha_{ppp}/D_{ppp}) - \beta_{ppp} \text{asinh}(\beta_{ppp}/D_{ppp}) - \gamma_{ppp} \text{asinh}(\gamma_{ppp}/D_{ppp}) + \delta_{ppp} \text{asinh}(\delta_{ppp}/D_{ppp}) - \sqrt{\alpha_{ppp}^2 + D_{ppp}^2} + \sqrt{\beta_{ppp}^2 + D_{ppp}^2} + \sqrt{\gamma_{ppp}^2 + D_{ppp}^2} - \sqrt{\delta_{ppp}^2 + D_{ppp}^2}); \]
\[ \text{else} \]
\[ D_p = h; \]
\[ \alpha_p = l + \delta_p; \]
\[ \beta_p = l + \delta_p; \]
\[ \gamma_p = l + \delta_p; \]
\[ M_{W_s_p}(i,j) = \frac{\varepsilon_r}{2} (\alpha_p \text{asinh}(\alpha_p/D_p) - \beta_p \text{asinh}(\beta_p/D_p) - \gamma_p \text{asinh}(\gamma_p/D_p) + \delta_p \text{asinh}(\delta_p/D_p) - \sqrt{\alpha_p^2 + D_p^2} + \sqrt{\beta_p^2 + D_p^2} + \sqrt{\gamma_p^2 + D_p^2} - \sqrt{\delta_p^2 + D_p^2}); \]
\[ \text{end} \]
\[ \text{end} \]
\[ \text{end} \]
% mutual-inductance for missing horizontal wires.

for i=1:M
    delta=0;
    alpha=l+1+delta;
    beta=l+delta;
    gamma=l+delta;
    M_W_d_1(i)=er/2*(alpha*asinh(alpha/h)-beta*asinh(beta/h)-
                      gamma*asinh(gamma/h)+delta*asinh(delta/h)-
                      sqrt(alpha^2+h^2)+sqrt(beta^2+h^2)+sqrt(gamma^2+h^2)-sqrt(delta^2+h^2));
end

for i=1:M-1
    delta=0;
    alpha=l+1+delta;
    beta=l+delta;
    gamma=l+delta;
    M_W_d_2(i)=er/2*(alpha*asinh(alpha/L)-beta*asinh(beta/L)-
                      gamma*asinh(gamma/L)+delta*asinh(delta/L)-
                      sqrt(alpha^2+L^2)+sqrt(beta^2+L^2)+sqrt(gamma^2+L^2)-sqrt(delta^2+L^2));
end

% total mutual-inductance

if N==1
    L_mutual=(sum(M_vertical(:))+sum(M_L(:))+sum(M_W_s_pp(:))+sum(M_W_s_p(:))+sum(M_W_d_1(:))+sum(M_W_d_2(:)));
else
    L_mutual=(sum(M_vertical(:))+sum(M_L(:))+sum(M_W_s_pp(:))+sum(M_W_s_pp(:))+sum(M_W_s_p(:))+sum(M_W_d_1(:))+sum(M_W_d_2(:)));
end

% self inductance.

L_self=er*Tline*(log(2*Tline/(B+C))+0.5-0.00181);

% total inductance

L_total=(L_self+L_mutual);
APPENDIX B

MASK OF THE DEVICES

3D inductance

Mask #1: bottom metal layer

Mask #2: via layer for BST etching

Mask #3: via layer for post

Mask #4: top metal layer
Resonator

Mask #1: bottom metal layer

Mask #2: via layer for BST etching

Mask #3: via layer for post

Mask #4: top metal layer
Tunable Inductor using VO$_2$ as Cantilevers

Mask #1: bottom metal layer

Mask #2: via layer

Mask #3: top metal layer

Mask #4: SiO$_2$ anchor

Mask #5: SiO$_2$ base

Mask #6: VO$_2$ etching layer

Mask #7: heater layer

Mask #8: Nickel layer
Tunable Inductor using VO$_2$ as Switches

Mask #1: bottom metal layer

Mask #2: via layer for BST etching

Mask #3: VO$_2$ etching layer

Mask #4: top metal layer