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**FABRICATION AND CHARACTERIZATION OF MICROWAVE
THIN-FILM MIM CAPACITORS**

Hien Do Ky, B. Eng.

**A thesis submitted in partial fulfillment of the requirements for
the degree of Master of Engineering at Carleton University**

**Ottawa-Carleton Institute of Electrical Engineering
Department of Electronics
Faculty of Engineering
Carleton University
Ottawa, Canada
September 1991
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FILM MIM CAPACITORS"**

submitted by Hien Do Ky in partial fulfillment of the requirements for the
degree of Master of Engineering

Thesis Supervisor
Chairman, Dept. of Electronics

Carleton University
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ABSTRACT

Metal-Insulator-Metal (MIM) Si_3N_4 capacitors have been fully characterized in the 0.5 GHz - 40 GHz frequency range. DC measurement, automatic network analyzer (ANA) calibration, S-parameter measurement, capacitor modelling and Q-factor measurement are described in detail.

The capacitors, ranging from 0.1 pF - 23 pF, are fabricated on 1 μinch surface finish 10 mil and 25 mil alumina substrates by sandwiching a 2000 Å - 5000 Å dielectric layer between a 2 μm - 4 μm electro-plated gold top layer and a 1 μm evaporated gold bottom layer. The DC measurements agree well with the designed values. The S-parameter measurements, in the 0.5 GHz - 26.5 GHz frequency range, are obtained by using the TRL (Thru-Reflect-Line) calibration method and are used to verify the physical modelling technique proposed here. Good agreement is found between the S-parameter measurements and the results derived from this optimize-free modelling technique.

A Q-factor measurement technique, termed the resonant frequency shift (RFS) technique, is employed in this study to determine the Q-values of microwave MIM capacitors in the extended 1 GHz - 40 GHz frequency range. The results of this study are compared to those derived from the previously constructed MIM capacitor physical models. The comparison shows that the the physical models, which are based solely on S-parameters, are inadequate in predicting MIM capacitor Q-values, as predicted. The delicate RFS technique is proven to be attractive for the measurement of Q-values of microwave MIM capacitors and other low-loss thin-film components.

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LIST OF SYMBOLS

a	incident travelling wave voltage
A	error box at input port of the automatic network analyzer (ANA)
β	microstrip gap coupling coefficient
b	reflected travelling voltage
B	error box at the output of the ANA
jB	susceptance
C_{ij}	capacitance from point i to point j
C_o	DC capacitance
$C_o(\epsilon_r)$	microstrip odd-mode capacitance
$C_e(\epsilon_r)$	microstrip even-mode capacitance
Δ	$S_{11} S_{22} - S_{12} S_{21}$
Δx	small increment in length
ϵ_{eff}	effective dielectric constant
ϵ_r	relative dielectric constant
ϵ_o	dielectric constant of free space=8.854 pF/m
e	voltage
f	frequency
Δf	3-dB bandwidth
Γ_i	reflection coefficient at port i
g	width of microstrip gap
G	conductance

h	substrate thickness
i	current
k_e, m_e	microstrip even-mode parameters
k_o, m_o	microstrip odd-mode parameters
K	correction factor for microstrip resistance calculation
K_g	correction factor for microstrip inductance calculation
λ	wave length
l_i	length of element i
L_o	DC inductance
L_i	self inductance of element i
L_{ij}	mutual inductance between i and j
λ	MANN mask generator resolution
m	magnitude of measured S_{21}
n	order of resonance
η	impedance ratio
P_{sys}	power dissipated by a resonator perturbed by a DUT
Q	quality factor
Q_i	quality factor of element i
Q_{sys}	quality factor of a resonator perturbed by a DUT
R_i	resistance of element i
R_a	2x2 cascading matrix of error box A
R_b	2x2 cascading matrix of error box B

R_d	2x2 cascading matrix of the line (delay) measurement setup
R_l	2x2 cascading matrix of the line (delay) itself
R_t	2x2 cascading matrix of the through measurement setup
S_{ii}	S-parameter ii
SiO_2	silicon dioxide
Si_3N_4	silicon nitride
t	thickness
T	$R_d R_t^{-1}$
U_i	time averaged energy stored in element i
U_E	time averaged energy stored in the electric field
U_H	time averaged energy stored in the magnetic field
U_{sys}	total time averaged energy stored in a resonator perturbed by a DUT
ϕ, θ	electrical lengths
v	voltage
v_{ph}	phase velocity in the transmission medium
v_{air}	velocity of light in air
ω	angular frequency
w	width
x	direction of propagation
X_i	reactance of element i
Y	admittance
Z_i	characteristic impedance of element i

LIST OF ABBREVIATIONS

A	ampere
Å	angstrom
ANA	automatic network analyzer
BW	bandwidth
CAD	computer-aided design
CAE	computer-aided engineering
CIF	Caltech intermediate format
CPU	central processing unit
CRC	Communications Research Centre
CVD	chemical vapour deposition
dB	decibel
DC	direct current
DUT	device under test
GHz	giga-Hertz
HMIC	hybrid microwave integrated circuit
J	joules
kΩ	kilo-ohm
LRL	Line-Reflect-Line
LRM	Line-Reflect-Match
μm	micro-metre
μinch	micro-inch

m	metre
MANN	David Mann mask generator's binary data format
MHMIC	miniature hybrid microwave integrated circuit
MHz	mega-Hertz
mil	mili-inch
MiM	metal-insulator-metal
MMIC	monolithic microwave integrated circuit
nH	nano-Henry
pF	pico-Farad
RF	radio frequency
RFS	resonant frequency shift
SOLT	Short-Open-Load-Thru
TAN	Thru-Attenuate-Network
TAR	Thru-Attenuate-Reflect
TFC	thin-film capacitor
TRL	Thru-Reflect-Line
TSD	Thru-Short-Delay
Ω	ohm
Ω /square	ohm per square

2.3 Design of Thin-Film MIM Capacitors

Thin-film MIM capacitors are designed with the knowledge of the following parameters:

- capacitance value
- dielectric relative permittivity and thickness
- capacitor and interconnection geometries
- fabrication tolerances

A complete model of thin-film MIM capacitors will be developed in Section 3.4.3. In this section, only one dominant element of that model will be discussed. It is the capacitance of the parallel-plate capacitor structure and is commonly described by Equation 2.1.

$$C = \frac{\epsilon_r \epsilon_0}{t} WL = C_A WL \quad (2.1)$$

Where:

- ϵ_0 is free space permittivity
- ϵ_r is dielectric relative permittivity
- W is capacitor width
- L is capacitor length
- t is dielectric thickness
- C_A is the capacitance per unit area

The value of C_A is process dependent, therefore, it must be provided to the designer by the fabrication staff. The area of the top electrode is first computed using the capacitance value (C) and capacitance per unit area (C_A) information. The values of W and L are then calculated using the geometry information of the MIM capacitor and its interconnections. It must be noted that a square geometry is often used because it offers lowest series resistance loss. However, in some applications, a rectangular geometry can certainly be employed to eliminate transitional discontinuities and parasitics. The bottom electrode dimensions are those of the top electrode plus the fabrication tolerances. The dimensions of the dielectric layer are further enlarged by adding the fabrication tolerances once more to the bottom electrode dimensions. The air bridge connecting the top electrode of the MIM capacitor to an output microstrip line must be designed within the allowable width-to-span ratio. If there is available space, two or more air bridges can be used to reduce transmission loss.

as well as discrete devices commercially available, especially in low noise and high power applications. MHMICs, on the other hand, have the flexibility of choosing the most suitable active elements for a particular application, yet maintain almost the same level of integration as MMICs. Additionally, MHMICs are more economical in medium volume applications because of lower cost and faster turn-around time. This is why research or industrial organizations and academic institutions often use MHMICs to complement and, in some cases, to replace MMICs. In the future, once MMIC technology has matured and gained more popularity, it is expected that the commercial market will concentrate more on this approach. However, MHMICs will always complement MMICs either to meet the cost-effectiveness and reliability standards of production engineering, or to meet specifications which can not be met solely by MMIC such as high power or low noise circuits. Also, high investment required in establishing and maintaining a MMIC facility is a major factor which will attract medium and small size companies to the MHMIC technology.

Currently, the multi-chip module concept is widely adopted by the microwave industries [3], [4], [5]. A typical module, or subsystem, can contain many small, moderate gain MMICs interminged with MHMICs to accommodate various application-specific requirements. Such a module combines the best of both worlds because it utilizes the potentially low-cost MMICs and the high-performance MHMICs simultaneously.

1.2 Thin-film Fabrication

In order to provide research facilities at Carleton University for studies on MHMICs, multi-chip modules and optical integrated devices, it was decided that an in-house thin-film fabrication process would be developed.

With its existing silicon facilities, CAD and automatic measurement systems, Carleton University was able to adapt to thin-film fabrication immediately [8]. For instance, metallizations of Ta, Cr, Au... are achieved using sputtering, and dielectric growth of SiO_2 is realized using the silane-based chemical vapour deposition (CVD) process. Also, layout packages, MIC simulation software packages are already supported on IBM PC's or VAX and SUN terminals. HP network analyzers (8409, 8720...), spectrum analyzers, noise figure and power meters... are available in both graduate and undergraduate laboratories.

The results of this study will be directly applicable to the development of a reliable and economical thin-film foundry for HMICs and MHMICs at Carleton University.

1.3 Thesis Objectives

This thesis studies three inter-related topics. They are the design, fabrication and characterization of microwave MIM capacitors on alumina substrates. Chapter 2 describes the fabrication and design procedures of microwave thin-film MIM capacitors. Theories involving CAD modelling approaches, optimization methods and experimental characterization of MIM capacitors are presented in Chapter 3. Chapter 4 explains the Resonant Frequency Shift (RFS) technique used to measure the Q-factor of thin-film lumped elements at microwave frequencies. In Chapter 5, characterization and Q-measurement results are compared against theoretical analyses given in the previous chapters. Finally, concluding remarks and recommendations for future research are given in Chapter 6. Also, computer simulation and layout descriptions, MIM capacitor fabrication steps, layout definitions, modelling procedures and Q-factor calculations are attached in Appendices I, II, III, IV and V, respectively.

CHAPTER 2: DESIGN AND FABRICATION OF MICROWAVE THIN-FILM MIM CAPACITORS

2.1 Introduction

The design of a thin-film MIC, in general, or a thin-film MIM capacitor, in specific, requires accurate and sophisticated device models and CAD systems in order to account for parasitic effects in elementary components and parasitic effects which arise from circuit layout [9], [10], [11]. Such effects are significant at microwave frequencies. In general, the thin-film circuit design procedure involves the following four basic steps:

- Analysis and simulation,
- Layout,
- Post-layout simulation,
- Mask generation.

The first step provides the designer with the circuit's physical parameters and nodal connections which produce a response meeting a predetermined set of specifications. Another CAD package then utilizes this information to produce a multi-layer drawing of the functional circuit. Post-layout simulation of the circuit is performed to account for the parasitics and couplings which are created from layout restrictions and space limitation. This is the simulation step which includes a more complete description of the circuit than that described in the first step. The layout may be modified if post-layout simulation shows undesirable outcomes. Steps two and three may be repeated as many times as necessary. From the final layout drawing, binary codes are created for the photolithography process in which the pattern generator is instructed to produce the masks. It is important to make sure that output data formats from all steps are fully compatible. If they are not, problems such as dimension round-off and coordinate shifting may occur and cause the photolithographic patterns, on the masks, to become disjoint or disfigured. Detailed descriptions of the analysis, simulation and layout tools are provided in Appendix I, hence, will not be further discussed here. In order to provide a complete reference, Appendix I also includes all necessary computer data transferal procedures among various data formats.

In this chapter, the design and fabrication of thin-film MIM capacitors are discussed. In Section 2.2, the fabrication of thin-film MIM capacitors in a typical MHMIC process is described. Section 2.3 provides rules and guidelines to be followed in the design and layout of MIM capacitors in order to satisfy

these thin-film MHMIC fabrication requirements. Finally, a summary of the chapter is given in Section 2.4.

2.2 Fabrication of Thin-Film MIM Capacitors

The MHMIC process is a multilayer fabrication technology [8]. Its layers are as described in Table 2.1.

Table 2.1

Layer Definitions of Thin-film Microwave MHMIC Process

DESCRIPTION	NAME	LAYOUT COLOUR
Resistive pattern	RES	Amber
Conductor	CON	Solid yellow
Bottom electrode window	CAP	Outlined yellow
Dielectric pattern	DIE	Sky blue
Top electrode or airbridge pedestal	PED	Navy blue
Airbridge	AIR	Green

The complete listings of all layer definitions can be found in the Microwave _ IC. BAT file, in Appendix II.

A thin-film MIM capacitor in the MHMIC process is basically a miniature parallel-plate overlay capacitor. It is composed of a thin bottom electrode, a dielectric layer and a thick top electrode. The abbreviation MIM arises from the metal-insulator-metal connection of the structure. The capacitance of a MIM capacitor is directly proportional to its effective area and the relative permittivity of the dielectric material used. However, it is inversely proportional to the thickness of the dielectric layer [10], [11]. The two electrodes are usually made from gold. The bottom electrode can be defined separately or as a part of the conductor pattern, depending on the choice of fabrication procedure. The latter is recommended because it will save one additional mask and fabrication step thus will be more cost effective. In Table 2.1, however, the two steps are listed separately to allow the designer the freedom to decide the appropriate choice for his or her fabrication process.

In this section, a brief description of the fabrication of a MIM capacitor and its interconnections are given to illustrate some of the steps of the MHMIC process, as shown in Figure 2.1. Reference [8] is to be consulted if a full description of the MHMIC process is required.

Figure 2.1 Typical MIM capacitor fabrication steps

As shown in Figure 2.1, a substrate, usually 10-25 mil polished alumina or quartz, is first selected and cleaned (degreased) to be used as a supporting surface for the MHMIC process (step (a)). Typical surface roughness of the selected substrate must be 4 μ inch or better. In case of alumina, material purity of 99.9% is preferred in order to meet manufacturing allowable tolerances.

A sputtered TiW-Au conductor pattern is then defined on the substrate using lift-off or etch-off technique (step (b)). The former is preferred because it can define patterns with features as fine as 5 μ m wide. The bottom electrode of the MIM capacitor is usually included in the conductor pattern. The thickness of this layer is about 1 μ m. It is to be noted that the 300 Å-TiW layer is acting as a adhesion layer (seed layer) between the alumina substrate and the gold (Au) metal only. Because of its high bulk resistivity, the TiW layer is always kept much thinner than the Au layer in order to maintain low loss. On the substrate, the areas at which the Au layer does not cover the TiW layer represent the definitions of thin-film resistors.

In the next fabrication step, a thin dielectric layer is deposited (evaporated or chemical vapour deposited) on top of the entire substrate (step (c)). Except for the areas defining the MIM capacitors or protecting the TiW resistors, the rest is etched off. At Carleton University, SiO₂ is used as the dielectric material whereas at the Communications Research Centre Si₃N₄ is used. The choice of dielectric material relies mainly on its availability. However, if a facility is to be constructed to support a MHMIC process, it must accommodate the one with higher relative permittivity, which is Si₃N₄ in this case. For a constant capacitance value and dielectric thickness, the higher relative permittivity will result in a smaller MIM area which will further facilitate the miniaturization purpose. The thickness of the dielectric layer usually ranges from 2000 Å to 5000 Å. The lower limit of this range comes from pinhole-free fabrication and voltage breakdown constraints. On the other hand, the upper limit comes from thermally induced cracking and miniaturization constraints.

The top electrodes of the MIM capacitor and the air bridge pedestals (for MIM capacitors, spiral inductors and cross-overs) are defined simultaneously on a thick layer of photoresist (step (d)). The entire substrate is then sputtered with a thin TiW-Au layer. The defined (exposed) areas on the photoresist allow the newly deposited metal layer to connect to either the tops of the

dielectric layer or the pedestal windows. At this stage, another photoresist layer with thickness slightly higher than the desired top electrode thickness is deposited on the substrate (step (e)). Again, the areas at which the air bridges are to be realized are defined. The exposed areas are then electroplated in a gold composite solution to the desired thickness. It is important to note that the electroplated gold layer must be kept thinner than the top layer of photoresist. In practice, the top electroplated-gold electrode is about 4 μm in thickness. Both the top and bottom photoresist layers are at the end chemically washed away to leave behind the air bridge structures (step (f)).

Finally, if a connection to the ground plane is required, a via hole can be drilled into the substrate (step (g)). Before drilling, the front surface of the substrate must be protected by a thick layer of photoresist. The back side of the substrate is sputtered and electroplated to build up a gold layer about 4 μm thick. The protecting layer of photoresist at the front can then be removed.

Typically, the top view of a MIM capacitor is as shown in Figure 2.2. It must be noted that the bottom electrode is completely covered by the dielectric layer to keep the two electrodes from shorting together. The top electrode is defined slightly smaller than the bottom electrode to help prevent the shortage problem. Hence, it is the dimensions of the top electrode which will determine the effective area and, ultimately, the capacitance of the MIM structure. It is also important to note that the air bridge shown in Figure 2.2 must have a span-to-width ratio of less than 5:1 in order to be able to support its own weight.

Figure 2.2 Top view of a MIM capacitor

2.3 Design of Thin-Film MIM Capacitors

Thin-film MIM capacitors are designed with the knowledge of the following parameters:

- capacitance value
- dielectric relative permittivity and thickness
- capacitor and interconnection geometries
- fabrication tolerances

A complete model of thin-film MIM capacitors will be developed in Section 3.4.3. In this section, only one dominant element of that model will be discussed. It is the capacitance of the parallel-plate capacitor structure and is commonly described by Equation 2.1.

$$C = \frac{\epsilon_r \epsilon_0}{t} WL = C_A WL \quad (2.1)$$

Where: ϵ_0 is free space permittivity
 ϵ_r is dielectric relative permittivity
 W is capacitor width
 L is capacitor length
 t is dielectric thickness
 C_A is the capacitance per unit area

The value of C_A is process dependent, therefore, it must be provided to the designer by the fabrication staff. The area of the top electrode is first computed using the capacitance value (C) and capacitance per unit area (C_A) information. The values of W and L are then calculated using the geometry information of the MIM capacitor and its interconnections. It must be noted that a square geometry is often used because it offers lowest series resistance loss. However, in some applications, a rectangular geometry can certainly be employed to eliminate transitional discontinuities and parasitics. The bottom electrode dimensions are those of the top electrode plus the fabrication tolerances. The dimensions of the dielectric layer are further enlarged by adding the fabrication tolerances once more to the bottom electrode dimensions. The air bridge connecting the top electrode of the MIM capacitor to an output microstrip line must be designed within the allowable width-to-span ratio. If there is available space, two or more air bridges can be used to reduce transmission loss.

Once all dimensions of a MIM capacitor are calculated, the physical model of that capacitor can be computed using the procedure proposed in Section 3.4.3.

2.7 Summary

The fabrication and design of a microwave thin film MIM capacitor have been described. All fabrication steps concerning the construction of an overlay capacitor were given. These included defining the bottom electrode, depositing the dielectric layer, defining the top electrode and constructing the air bridge. The bottom electrode is 1 μm sputtered TiW-Au and defined using the lift-off technique. The dielectric layer is 2000-5000 Å SiO_2 or Si_3N_4 . The top electrode and the air bridge are 4 μm electroplated Au and are defined using a double coverage process.

The design of a MIM capacitor was also described. The dimensions of the top electrode were determined from the desired capacitance value, capacitance per unit area and the geometry of the capacitor. The dimensions of the bottom electrode were larger than those of the top electrode by the tolerances given by the fabrication. The dimensions of the dielectric layer were made to cover the bottom electrode entirely to prevent shorting the two electrodes.

CHAPTER 3: MODELLING OF THIN-FILM OVERLAY CAPACITORS

3.1 Introduction

The development of an MHMIC technology requires a complete and accurate characterization of all its lumped elements, in an acceptable range of microwave frequencies. Electrical models are developed to simulate correctly both DC and RF responses.

Microwave lumped elements, which must be much smaller than the wavelengths at the frequencies of interest, exhibit low-loss and broadband responses. Hence, delicate techniques must be devised to ensure the accuracy and repeatability of all measurements. Then the electrical models can be verified against these measurements to prove their validity.

There are three different techniques for measuring resistance and reactance of a lumped element. They include:

- i) Direct impedance measurement using the automatic network analyzer or some other broadband impedance measurement instrument
- ii) The resonant frequency shift (RFS) technique
- iii) The study of the reactance or susceptance characteristics near the self resonance of lumped element under test or of a combination of this element with another.

In this chapter, the direct measurement method using the HP8510 automatic network analyzer is examined. The MIM capacitors, on alumina and quartz, are probed up to 26.5 GHz. Two-port measurements have been carried out using the TRL (Through-Reflect-Line) calibration technique and a Cascade Microtech wafer prober set-up.

The modelling technique applied in this thesis is termed physical modelling. Model parameters are computed primarily from the dimensions and geometry of a capacitor of interest. The computations involve treating each discontinuity in the geometry separately and then electrically combining all discontinuities together at the final stage. The term discontinuity used here must be understood in the most general sense. Any basic microstrip structure situated between the input and output transmission lines of the MIM capacitor must be treated as a discontinuity, even the parallel plate structure. This modelling technique is compared against other techniques and the measurements to confirm its performance.

The RFS technique is discussed and implemented in the next chapter, while the third method is left unexplored in this thesis.

3.2 DC Measurements

The DC capacitances of the MIM capacitors are measured at 1 MHz using a HP4280 C-V meter. These measurements are necessary to identify capacitors which may have problems associated with pin holes and/or air-bridges before proceeding with the RF evaluations.

The test set-up is as shown in Figure 3.1.

Figure 3.1 C-V Measurement Set-up.

3.3 Automatic Microwave Measurements

3.3.1 TRL Calibration

Conventionally, an automatic network analyzer (ANA) is comprised of two 4-port reflectometers as shown as shown in Figure 3.2.

Figure 3.2 Conventional automatic network analyzer

The nonideal reflectometer can be modelled by an ideal four-port cascaded with a two-port "error-box", as shown in Figure 3.3.

Figure 3.3 Error model for S-parameter measurement using an automatic network analyzer

A six-port reflectometer can also be modelled in a similar manner, as suggested by Egen and Susman [14], [15].

It can be seen from Figure 3.3 that the parameters of the error boxes A and B are to be determined by the ANA calibration. Once characterized, these error boxes can be effectively removed to provide an "exact" determination of the signals at the measurement planes.

One calibration technique which can be used here is the "through-reflect-line" (TRL) procedure [16]. Other techniques such as "through-short-delay" (TSD), "short-open-load-through" (SOLT), "Line-reflect-match" (LRM), "line-reflect-line" (LRL), "Through-attenuate-reflect" (TAR), "Through-attenuate-network" (TAN) [17], [18], [19], [20] can also be used but will not be discussed here.

As the name describes, the TRL calibration technique measures three connection combinations of error boxes A and B in a through, a reflective and a delay configuration. The term "line" is used in place of "delay" because it is more descriptive of the TRL technique. This line length is arbitrary and can be unknown, but must be different from $\lambda/2$. Also, it need not be free of dispersion.

In Figure 3.3, the S-parameters measured between the fictitious detector planes are:

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \quad (3.1)$$

Rearranging (3.1) gives,

$$\begin{bmatrix} b_1 \\ a_1 \end{bmatrix} = \frac{1}{S_{21}} \begin{bmatrix} -\Delta & S_{11} \\ -S_{22} & 1 \end{bmatrix} \begin{bmatrix} a_2 \\ b_2 \end{bmatrix} \quad (3.2)$$

$$= R \begin{bmatrix} a_2 \\ b_2 \end{bmatrix}$$

where $\Delta = S_{11} S_{22} - S_{12} S_{21}$
and the wave cascading matrix R is,

$$R = \frac{1}{S_{21}} \begin{bmatrix} -\Delta & S_{11} \\ -S_{22} & 1 \end{bmatrix} \quad (3.3)$$

It is important to note that the R matrix for two or more 2-ports in cascade is merely the product of the individual matrices.

Let the cascading matrices of the error boxes A and B be denoted by R_a and R_b , respectively. Then each connection can be consider in turn as follows:

a) In the through connection,

$$R_t = R_a R_b \Rightarrow R_b = R_a^{-1} R_t \quad (3.4)$$

where R_t is the cascading "through" matrix.

b) In the line combination,

$$R_d = R_a R_l R_b \quad (3.5)$$

where:

R_d is the cascading "line" matrix

R_l is the matrix which represents the inserted line.

Substitution of (3.4) into (3.5) gives:

$$T R_a = R_a R_l \quad (3.6)$$

Where $T = R_d R_t^{-1}$.

Clearly, T can be found from the through and the line measurement.

c) In the reflect connection,

with an unknown element connected, which generates some nonzero reflection, the Γ 's at the measurement planes can be obtained. Let these be Γ_1 and Γ_2 respectively.

Using (3.1), it can be shown that:

$$\begin{aligned} \Gamma_2 S_{11} + \Gamma_1 S_{22} - \Delta &= \Gamma_1 \Gamma_2 & (3.7) \\ \text{where } \Delta &= S_{11} S_{22} - S_{12} S_{21} \\ \Gamma_1 &= \frac{b_1}{a_1} \\ \Gamma_2 &= \frac{b_2}{a_2} \end{aligned}$$

From all three measurements, 4 equations are found (3.4-3.7). These are enough to solve for the 4 unknown complex pairs of parameters of R_a . Once R_a is found, R_b can be calculated from equation (3.1). The detailed calculation of R_a is given by Egen and Hoer [16].

With R_a and R_b determined, the ANA is effectively calibrated and ready for measurements. At CRC, a Turbo Pascal program written by S. Meszaros and the controlling software of the HP 8510 are currently used to perform TRL de-embedding and calibration, respectively.

3.3.2 Microwave Coplanar Wafer Prober

Microwave coplanar wafer probes have been reported to function satisfactorily up to 50 GHz [21], [22]. In this work a Cascade Microtech prober is used in conjunction with the HP8510 network analyzer to wafer-evaluate both the MHMIC and the MMIC technologies. Calibration of the measurement system can be performed quickly and directly at the probe tips. The TRL calibration procedure is used here, even though the LRM technique is also recommended by Cascade Microtech, to de-embed the systematic errors in the cables, the connectors and the probes.

A set of coplanar impedance "standards" consisting of open, short, load and through line has been developed by Cascade Microtech for the purpose of de-embedding.

Figure 3.4 Microwave coplanar wafer prober setup

(a) Through or delay calibration standard

(b) Offset short calibration standard

Two different lengths of delay line (one very short and the other longer) and an offset short are used as calibration standards. A delay line and an offset short for the calibration process are illustrated in Figure 3.4. The TRL calibration routine is already programmed directly into the HP8510 for real-time de-embedding purposes.

3.4 Modelling of MIM capacitors

3.4.1 Distributed Approaches

3.4.1.1 Open Circuited Transmission Lines

Ingalls and Kent [23] have suggested an open-circuited transmission line model for high-Q multilayer capacitors, based on the measured characteristics of these elements. A series-connected capacitor exhibits periodic resonances, similar to those of an open-circuited transmission line, as shown in Figure 3.5. The dashed envelopes shown in this figure indicate the upper and lower theoretical limits of the troughs and peaks of the transmission response.

Figure 3.5 Typical transmission response $|S_{21}|$ of a series thin-film capacitor

From this observation, it seems appropriate to model multilayer capacitors using open-circuited transmission lines. The model developed by Ingalls and Kent [24] for each periodic section of a multilayer capacitor is illustrated in Figure 3.6. In this figure, l_2 is the actual length of the multilayer capacitor, while l_1 is the transmission line length of half of a folded section. Theoretically, the sum of l_1 and l_0 must yield the length l_2 . However, experimental results [23] indicated otherwise.

In this scheme, a transmission line, with length $l_1/2$, and characteristic impedance Z_1 , is to be computed to replace the half of the folded section in one period. Once one period is modelled, the entire multilayer capacitor can

be realized by connecting these models together in parallel. The result is a periodically loaded transmission line.

Figure 3.6 Model of one periodic section of a multilayer capacitor [24].

- (a) Plan view
- (b) Cross-section through AA' in (a)
- (c) Lumped equivalent circuit of one period
- (d) Transmission line equivalent circuit of one period

l_1 and Z_1 are related by:

$$l_0 + 2\eta l_1 = Z_0 2\eta v_{ph} C_0 \quad (3.8)$$

$$l_0 + \left(\frac{2}{\eta}\right) l_1 = \frac{v_{ph}}{Z_0} L_0 \quad (3.9)$$

$$\eta = \frac{Z_0}{Z_1} \quad (3.10)$$

where

η = impedance ratio of line l_0 and l_1

Z_0 = characteristic impedance of line l_0

v_{ph} = phase velocity in the transmission medium

C_0 = DC capacitance

L_0 = DC inductance (with capacitor shorted)

l_0 , l_2 and w are as shown in Figure 3.6

From (3.8) and (3.9), l_1 and Z_1 can be calculated if C_0 and L_0 are known. In order to measure L_0 , the multilayer capacitor must be dissected. In practice, this is very difficult to accomplish. In addition, because of measurement difficulties, values of C_0 and L_0 are usually measured with less than adequate precision. Hence, the derivation of the model is not very repeatable.

Another approach is to optimize l_1 and η against measured S-parameter data, then calculate Z_1 using (3.10). The resulting values of l_1 and Z_1 may not, however, have any direct relationship to the capacitor's geometry. Furthermore, the quality of the fit may have no relevance to the validity of the equations proposed by this technique.

3.4.1.2 R, L, C per Unit Length

Mondal [25], [26] has suggested a distributed-circuit model for a MIM capacitor based on a broadside-coupled transmission line approach. For a typical MIM capacitor of Figure 3.7a), the model proposed by Mondal for a short length Δx of the parallel plate region is as shown in Figure 3.7b).

Figure 3.7 Distributed circuit model of MIM capacitor [25]

(a) Cross-section of a MIM capacitor

(b) Equivalent circuit for a unit length Δx

Where:

- L_{12} is the mutual inductance per length
- L_{11} , L_{22} are the self inductances per length
- C_{10} , C_{20} are the coupling capacitances per length between the electrodes and the ground plane
- C_{12} is the mutual capacitance per length between the electrodes
- R_1 , R_2 , G are the series and shunt positive losses

From Figure 3.7, the coupled-mode transmission line equations for propagation in the x-direction are:

$$\frac{-dv_1(x)}{dx} = (R_1 + j\omega L_{11}) i_1(x) - j\omega L_{12} i_2(x) \quad (3.11)$$

$$\frac{-dv_2(x)}{dx} = (R_2 + j\omega L_{22}) i_2(x) - j\omega L_{12} i_1(x) \quad (3.12)$$

$$\frac{-di_1(x)}{dx} = (G + j\omega(C_{10} + C_{12})) v_1(x) - (G + j\omega C_{12}) v_2(x) \quad (3.13)$$

$$\frac{-di_2(x)}{dx} = (G + j\omega(C_{20} + C_{12})) v_2(x) - (G + j\omega C_{12}) v_1(x) \quad (3.14)$$

If the length of the capacitor is l then the boundary conditions for the capacitor in Fig. 3.7 are:

$$i_1(x=l) = 0: \text{ no current at the open end of the top electrode} \quad (3.15)$$

$$i_2(x=0) = 0: \text{ no current at the open end of the bottom electrode} \quad (3.16)$$

Applying these two conditions to (3.11), (3.12), (3.13) and (3.14), the inductance parameters of the model are related to the capacitance parameters by:

$$\begin{bmatrix} L_{11} & L_{12} \\ L_{21} & L_{22} \end{bmatrix} = \frac{1}{v_{\text{air}}^2} \begin{bmatrix} C_{10} + C_{12} & -C_{12} \\ -C_{12} & C_{20} + C_{12} \end{bmatrix}_{\text{air}}^{-1} \quad (3.17)$$

where

v_{air} = velocity of light in air

C_{10} , C_{20} and C_{12} are calculated with $\epsilon_r = 1$ (air)

The mutual capacitance C_{12} is calculated by:

$$C_{12} = \frac{\epsilon_r \epsilon_0}{t} l \quad [\text{pF/m}] \quad (3.18)$$

where

C_{12} = capacitance of the parallel plate structure per unit width

ϵ_r = relative dielectric constant

$\epsilon_0 = 8.854 \text{ pF/m}$

t = dielectric thickness

l = length of top electrode

The self-capacitances C_{10} and C_{20} are given by:

$$C_{20} = C_{\text{shunt}} + (C_{\text{stripline}} - C_{\text{shunt}}) \frac{\epsilon_{\text{eff}}}{\epsilon_r} \quad [\text{pF/m}] \quad (3.19)$$

$$C_{10} = C_{\text{total}} - C_{20} \quad [\text{pF/m}] \quad (3.20)$$

where

$$C_{\text{shunt}} = \epsilon_0 \epsilon_r$$

h = substrate thickness

$$C_{\text{stripline}} = \frac{1}{2} \left[\frac{Z_{\text{stp}} \vartheta_{\text{air}}}{\sqrt{\epsilon_r}} \right]^{-1} \quad [\text{pF/m}]$$

Z_{stp} = characteristics impedance of a stripline with the same width as the top electrode of the MIM capacitor.

$$C_{\text{total}} = \left[\frac{Z_0 \vartheta_{\text{air}}}{\sqrt{\epsilon_r}} \right]^{-1} \quad [\text{pF/m}]$$

Z_0 = characteristic impedance of a microstrip line with the same width as the top electrode of the MIM capacitors.

In this modelling scheme, the elements C_{12} , C_{10} and C_{20} are first calculated with the value of ϵ_r of interest (for example, $\epsilon_r = 7$ for Si_3N_4). Then the inductances are computed with a different set of capacitances, which are associated with $\epsilon_{r \text{ air}} = 1$, using (3.12). The resistance losses are those of the metallic electrodes, combined with an increment directly proportional to the square root of frequency. The calculated values of the elements are used as starting values to optimize the model's response to measured data.

This model shows good agreement with the measurements after its elements have been optimized. The equations derived here are applicable to microwave circuit simulation packages, however, they are not easily implemented.

3.4.2 Lumped Equivalent Circuit Generation By Computer

An interesting method of lumped equivalent circuit generation, devised by Baden Fuller et. al. [27]-[30], allows a computer program to modify the model as necessary. The originally derived equivalent circuit is optimized by this program to match the measured data over the entire frequency range of interest. The program includes a sequence of steps as follows:

- i) Optimization of the model's element values only;
- ii) Removal of elements and nodes within the existing topology to simplify the circuit without any loss of performance;
- iii) Addition of elements within the existing topology of the network;
- iv) Addition of new topographic nodes to the existing model. This allows the growth of quite complex structures from a relatively simple starting model. Furthermore, in this step, the program automatically tests for removal of elements following any addition of new elements or new nodes.

For example, a simple equivalent circuit for an overlay capacitor, as in Figure 3.8, can be computer expanded to those of Figures 3.9a) and b), respectively.

As shown, the program enters the element and node addition routines alternately until a circuit of the required accuracy or complexity has been obtained. The criterion for completion may be either an error function less than a specified quantity, or a maximum number of components or nodes allowed in the circuit. The program is designed to go through all its steps without any operator intervention. It makes all decisions about the suitability of a particular circuit solely on the basis of minimizing the error function. Thus, not all of the resulting circuits are realistic physical models, as the program is only seeking an improved fit to the measured data. However, data on each converged model are stored and presented at the output, so it is left to the user to identify the most appropriate circuit. Limited steering of the model evolution is possible by starting the program at a slightly changed intermediate topology, which may contain a more physically desirable structure than the later circuits. Evidently, despite their complexity, the equivalent circuits of Figure 3.9 showed no significant improvement in performances when compared to those of the circuit of Figure 3.8.

Figure 3.8 Simple equivalent circuit of an overlay capacitor

(a)

(b)

Figure 3.9 Computer generated models of capacitor in Figure 3.8 [27]-[30]

(a) One possible solution

(b) Another possible solution

3.4.3 Proposed Physical Modelling Approach

The afore-mentioned modelling techniques have their advantages and drawbacks. Similarly, they all suffer from mathematical and computational complications which hinder their wide-spread application. Moreover, they require some degree of computer optimization which, in most cases, dictates their final accuracy [23]-[33]. Such features make these techniques undesirable in defining in-house user models for MHMIC lumped elements.

A physical modelling method is proposed in this section to fulfill the above requirements. The advantages of this technique are:

- i) Accurate characterization of the MIM capacitors up to the first parallel resonance frequency;
- ii) Derivation of the model's parameters based solely on the geometry of the capacitor in consideration, but not on computer optimization;
- iii) Simple implementation of the model in existing microwave circuit simulations.

The model is based on the physical geometry of the MIM capacitor. For example, a series-connected capacitor and its complete equivalent circuit are depicted in Figures 3.10 a) and b), respectively.

Figure 3.10 Physical modelling technique

(a) Physical layout of a series-connected MIM capacitor

(b) Complete equivalent circuit for capacitor in (a)

The complete equivalent circuit is derived by modelling all discontinuities and coupling mechanisms, associated with the MIM capacitor, using lumped elements. Figure 3.10 b) shows the MIM capacitor as connections of a microstrip step at the input in series with a lossy transmission line (bottom electrode), a low-loss capacitor, another lossy transmission line (top electrode and airbridge), and an output microstrip step together with capacitive coupling between the input and output ports. The microstrip steps can be represented by C_i and C_o . The short lengths of transmissions lines (i.e. both electrodes) are series RL networks. Any inductive loss associated with the steps can be conveniently incorporated into these RL networks. The derived MIM capacitance is C_1 . The parasitic coupling mechanism between the input and output ports is represented by the π -network comprised of C_a and C_b .

The derivation of the physical lumped model involves calculations for each discontinuity, using appropriate closed-form expressions as follows.

a) Microstrip step

For a microstrip step with the geometry of Figure 3.11 a), its electrical equivalence is that of Figure 3.11 b).

Figure 3.11 Microstrip step

(a) Topology of a microstrip step

(b) Electrical model of a microstrip step

According to Gupta et al [34], the shunt capacitance introduced by the microstrip step is given by :

$$C_{\text{step}} = \begin{cases} \sqrt{w_1 w_2} [(4.386 \ln \epsilon_r + 2.33) \frac{w_2}{w_1} - 5.472 \ln \epsilon_r - 3.17], \\ \text{for } \epsilon_r \leq 10 \text{ and } 1.5 \leq \frac{w_2}{w_1} \leq 3.5 \\ \sqrt{w_1 w_2} [56.46 \ln \left(\frac{w_2}{w_1} \right) - 44], \\ \text{for } \epsilon_r = 9.6 \text{ and } 3.5 \leq \frac{w_2}{w_1} \leq 10 \end{cases} \quad (3.21)$$

The series inductance introduced by the microstrip step is:

$$L_{\text{step}} = h \left[40.5 \left(\frac{w_2}{w_1} - 1 \right) - 32.57 \ln \left(\frac{w_2}{w_1} \right) + 0.2 \left(\frac{w_2}{w_1} - 1 \right)^2 \right] \quad (3.22)$$

where h is the substrate height.

This type of discontinuity occurs at both the input and output ports.

b) Microstrip transmission line

For a microstrip line with a short length l , thickness t and width w , the equivalent series inductance is [35]:

$$L_{\text{line}} = 2 \times 10^{-4} l \left[\ln \left(\frac{1}{w+t} \right) + 1.193 + 0.2235 \left(\frac{w+t}{1} \right) \right] K_g \quad (3.23)$$

where the correction factor K_g is

$$K_g = 0.57 - 0.145 \ln \left[\frac{w}{l} \right]$$

Also, the equivalent series resistance is:

$$R_{\text{line}} = K \frac{R_s}{2} \frac{1}{(w+t)} \quad (3.24)$$

where K is the correction factor given by

$$K = 1.4 + 0.217 \ln \left(\frac{w}{5t} \right)$$

R_g is the sheet resistance of the thin-film conductor.

($R_g = 0.029 \Omega/\text{square}$ for TiW-Au film)

The total inductance, L , and resistance, R , in the model include contribution from both feedlines and both electrodes.

c) Microstrip gap

The main coupling between the input and output ports occurs underneath the airbridge. It is as shown in Figure 3.12.

Figure 3.12 Microstrip gap

(a) Microstrip gap in a MIM structure

(b) Electrical model of a microstrip gap

Figure 3.12 b) shows the electrical equivalent circuit of the microstrip gap with linewidth, w , gap spacing, g , and substrate height, h . The parameters in the equivalent circuit are computed by [34]:

$$C_a = \frac{1}{2} (C_o(\epsilon_r) - \frac{1}{2} (C_e(\epsilon_r)) \text{ [pF]} \quad (3.25)$$

$$C_b = \frac{1}{2} C_e(\epsilon_r) \text{ [pF]} \quad (3.26)$$

where

$$\begin{aligned} C_o(\epsilon_r) &= w \left(\frac{g}{w} \right)^{m_o} \exp(k_o) \left(\frac{\epsilon_r}{9.6} \right)^{0.8} \\ C_e(\epsilon_r) &= w \left(\frac{g}{w} \right)^{m_e} \exp(k_e) \left(\frac{\epsilon_r}{9.6} \right)^{0.9} \quad \text{for } 2.5 \leq \epsilon_r \leq 15 \\ m_o &= \frac{w}{h} \left[0.169 \log \left(\frac{w}{h} \right) - 0.3853 \right] \\ k_o &= 4.26 - 1.453 \log \left(\frac{w}{h} \right) \quad \text{for } 0.1 \leq \frac{g}{w} \leq 1.0 \end{aligned}$$

$$m_e = \begin{cases} 0.8675 & \text{for } 0.1 \leq \frac{g}{w} \leq 0.3 \\ \frac{1.565}{\left(\frac{w}{h}\right)^{0.16}} - 1 & \text{for } 0.3 \leq \frac{g}{w} \leq 1.0 \end{cases}$$

$$k_e = \begin{cases} 2.043 \left(\frac{w}{h}\right)^{0.12} & \text{for } 0.1 \leq \frac{g}{w} \leq 0.3 \\ 1.97 - \frac{0.03}{\left(\frac{w}{h}\right)} & \text{for } 0.3 \leq \frac{g}{w} \leq 1.0 \end{cases}$$

d) Parallel plate capacitance

From experimentation and suggestions given by other publications [25], [36], it was found that the capacitance of the MIM structure could be sufficiently calculated using the expression for a parallel plate capacitor with area determined by the upper electrode dimensions:

$$C = \frac{\epsilon_r \epsilon_0}{t} w_1 w_2 \quad (3.27)$$

where

ϵ_r = relative dielectric constant

ϵ_0 = 8.854 pF/m

t = dielectric thickness

w_1 = top electrode width

w_2 = top electrode length

The complete equivalent circuit, of Figure 3.10 b), is quite capable of characterizing the MIM capacitor with very good accuracy. However, a simplified version of this circuit may also be used for the same purpose, without much degradation in performance. If carefully manipulated, the complex complete equivalent circuit can be reduced to that of the lumped physical model proposed in Figure 3.13.

Figure 3.13 Proposed physical model for a MIM capacitor

The complete equivalent circuit of the MIM capacitor can be simplified by several methods. One obvious choice is to equate the Z matrices of the circuits of Figure 3.10 b) and Figure 3.13 and then derive each element, of the circuit in Figure 3.13, accordingly. This approach, however, is too mathematically involved and computer-time consuming, hence, will not be further pursued.

The second approach is to consider the distribution of the coupling effects of the parallel plate structure. The couplings, through the dielectric (C_1) and through the substrate (C_2), can be considered to take place entirely within the MIM structure [37]. Also, most inductive and resistive losses occur due to the bottom electrode, the top electrode and the airbridge only. The inductances introduced by the microstrip steps at both ports are small, hence, they can be summed directly to the overall inductive loss to account for the worst-case consideration. The shunt capacitances at the input, or output, port can be similarly combined. The resulting equivalent circuit is that of Figure 3.13.

The model will consist of:

$$L_{\text{total}} = L = L_{\text{bottom electrode}} + L_{\text{top electrode}} + L_{\text{airbridge}}$$

$$R_{\text{total}} = R = R_{\text{bottom electrode}} + R_{\text{top electrode}} + R_{\text{airbridge}}$$

$$C_{\text{total}} = C_1 = C_{\text{parallel plate}}$$

$$C_{\text{coupling}} = C_2 = C_{\text{coupling due to microstrip gap}}$$

$$C_{\text{stray}} = C_3 \text{ or } C_4 = C_{\text{stray due to step}} + C_{\text{stray due to gap}}$$

3.5 Summary

In this chapter, all important steps needed to characterize thin-film MIM capacitors, and thin-film passive elements in general, were studied. First of all, the DC evaluation was considered to detect fabrication flaws such as pin holes or misalignments. Secondly, the RF evaluation which focused on the S-parameter measurements only was examined to validate the proposed physical lumped modelling technique. Q measurements of thin-film elements, to be discussed in the next chapter, will further confirm the integrity of the modelling technique.

Regarding microwave automatic measurements, the Through-Reflect-Line (TRL) calibration procedure and the Cascade-Microtech wafer probing technique were described in detail. In the future, for broadband measurements the Through-Attenuate-Network calibration technique is to be performed.

Also, various thin-film passive element modelling techniques were compared against the proposed modelling scheme. These are the distributed RLC and transmission line techniques. For the MHMIC technology, these techniques have proven to be impractical due to their computation intensive nature. The proposed physical modelling technique, on the other hand, is directly applicable to existing CAE and CAD software packages because its algorithm permits very fast simulation and optimization.

CHAPTER 4: MICROWAVE Q-MEASUREMENTS OF THIN-FILM ELEMENTS

4.1 Introduction

In Chapter 3, thin-film MIM capacitors were modelled based on their physical geometries. The results obtained, to be shown in the next chapter, will confirm the validity of the proposed model. To further confirm the model and to fully characterize thin-film MIM capacitors, their Q values must be determined. A technique, termed the resonant frequency shift (RFS), is described in this chapter to facilitate the Q-measurement requirements for low-loss thin-film elements in general, and for thin-film MIM capacitors in particular.

The elaborate RFS technique is required here because other common techniques such as calculations from S-parameters and Q-bridge measurements are incapable of determining very low losses accurately, especially resistive losses, at microwave frequencies [38], [39]. The difficulties with these latter two techniques arise from the imperfections of the instruments used. The actual DUT's loss is masked by the losses and uncertainties of the instruments. The RFS technique, on the other hand, determines the Q-value and DUT's reactance indirectly from changes which occur when a resonant structure is loaded by a DUT. The indirectness of this approach effectively removes the dependence of measurement accuracy from the instrument's uncertainties. The RFS technique, moreover, can be extended up to and beyond the K_a band because only the changes of the resonant properties, due to loading, are of interest here, not the actual resonant properties themselves, as long as the resonances can be observed.

In this chapter, an end-coupled linear resonant structure is analyzed to study its responses due to the perturbation of a foreign lumped element, under a sinusoidal excitation. From the measured results of loaded and unloaded Q-values and resonant frequencies of the resonator, the reactance, Q-values and loss of the DUT are calculated as functions of frequency [40], [41].

4.2 Quality Factor of a Resonant Structure

Microwave measurements of transmission line resonant structures reveal that resonances occur at harmonic multiples of a fundamental frequency. A typical transmission response of a resonant structure is that of a high Q and narrow band bandpass filter, as shown in Figure 4.1.

Figure 4.1 Typical transmission response of a transmission line resonant structure

Q factor is a defined quantity which describes the selectivity of the resonant structure. It is commonly defined as the ratio of the resonant frequency to the 3 dB bandwidth of the spectral response. As quantitatively depicted by Figure 4.1, only frequencies at multiples of f_0 are passed. The Q factor of the resonator degrades as frequency increases. This is clearly demonstrated by the broader responses of the resonance curves at higher order modes. As will be shown later, Q factor is inversely proportional to resistive loss which, in turn, is a direct function of the square root of frequency.

Usually, if only a particular resonance is of interest, a lumped model (using L, R and C) can be devised to describe such a phenomenon in a narrow band near resonance. The Q factor at the resonance of interest is then defined by:

$$Q = \frac{\text{Magnitude of reactance}}{\text{Resistance}} \quad (4.1)$$

However, the lumped model is not valid at other frequencies because it was specifically constructed to describe a particular resonance only. Thus, it is

best to characterize the resonant structure using a distributed model, per unit length, from which the Q factors of all resonances can be deduced. Also, using the distributed model, the resonance conditions and Q equation can be conveniently expressed in terms of stored energies, dissipated powers and currents or voltages associated with the resonant structure [42].

Let U_H and U_E be the time average of the magnetic and electric energies stored in the low-loss resonant structure, which is excited by a current i and voltage e , respectively. The input impedance and admittance of the resonator are described by:

$$Z(\omega) = \frac{j\omega^2 (U_H - U_E)}{0.5 ii^*} \quad (4.2)$$

$$Y(\omega) = \frac{j\omega^2 (U_E - U_H)}{0.5 ee^*} \quad (4.3)$$

where ω is frequency expressed in radians.

At frequencies for which $U_H=U_E$, extrema of power dissipation are obtained because either $Z(\omega)$ or $Y(\omega)$ diminishes. At $Z(\omega)=0$, a series resonance is found and the power dissipation by the resonator is zero. At $Y(\omega)=0$, a parallel resonance is found and the power dissipation is also zero. The Q factor of a resonance relates directly the ratio of the stored energy in the resonant structure over the power dissipated by the resonator. For the same stored energy U , the higher the dissipated power P , the poorer the frequency selectivity (lower Q value) and vice versa.

$$Q_{sys} = \omega \frac{U_{sys}}{P_{sys}} \quad (4.4)$$

Where the subscript sys denotes the entirety of the resonant structure, including all of its constituents.

If the resonant structure consists of N independent elements, the total stored energy and dissipated power are simply the sum of those of all elements. That is:

$$P_{sys} = \sum_{i=1}^N P_i = \sum_{i=1}^N \omega \frac{U_i}{Q_i} \quad (4.5)$$

$$U_{sys} = \sum_{i=1}^N U_i$$

Rewriting (4.4), one gets [41]:

$$\frac{U_{sys}}{Q_{sys}} = \frac{\sum_{i=1}^N U_i}{Q_{sys}} = \sum_{i=1}^N \frac{U_i}{Q_i} \quad (4.6)$$

Using (4.6), a Q value of an element j can be computed with the knowledge of Q_{sys} , U_i , U_j and Q_i , $i=1, \dots, N$, $i \neq j$.

4.3 Perturbation of a Resonant Structure

The theory derived in the previous section is directly applicable to the case of a perturbed resonator. When a resonator is perturbed by a DUT, it changes in characteristics to accommodate the intrusion. It is the difference in characteristics between the unloaded and loaded versions of the resonator that provides the vital information which allows the calculations of the DUT's Q value, reactance and total loss to be performed [43], [44]. Because the resonator exhibits resonant responses also at harmonic multiples of the fundamental frequency, the loading effects due to the DUT can be seen at several related frequencies, thus permitting a broadband characterization of the DUT.

In this study, a single-section end-coupled microstrip resonator is selected so that it can be fabricated also by the MHMIC process. Very loose couplings are required to ensure the highest available unloaded Q value. The resonator is fabricated on 10 mil alumina substrates with the DUT inserted, in series, at the center because the symmetry of the loaded system will result in significant simplifications of the theories involved.

Initially, the energy distribution between the resonator and the DUT, a thin-film MIM capacitor in this case, is examined. Then the resonance condition of a loaded resonator is established. The RFS theory is applied so that the reactance, Q value and resistive loss of the DUT can be calculated at relevant order resonances.

4.3.1 Energy Stored in a Single-Section End-Coupled Microstrip Resonator

Figure 4.2 depicts a single-section end-coupled microstrip resonator and its voltage and current standing wave pattern of the fundamental order mode, under sinusoidal excitation.

Figure 4.2 Unloaded resonator and its responses

(a) An unloaded single-section end-coupled linear resonator

(b) Fundamental mode current & voltage responses of resonator in (a)

The current is maximum at the center of the resonator and diminishes to zero at the open ends. This behaviour is typical for all odd-order modes. The even-order modes, on the other hand, exhibit a current zero at the center and at the open ends. This is why only odd-order resonances are affected by a series DUT connected at the center of the resonator [41]. For example, the standing wave patterns of the fundamental mode are altered by the insertion of a series capacitor at the center as shown in Figure 4.3. Similar effects will be experienced by all other odd-order modes.

Figure 4.3 Loaded resonator and its responses

(a) A loaded single-section end-coupled linear resonator

(b) Fundamental mode current & voltage responses of resonator in (a)

In general, the series connection of a DUT at the center affects all odd-order modes, whereas a shunt connection affects the even-order modes. Perturbations at other locations may affect all modes, which in theory would yield more information but very complicated to analyze.

Let the physical and electrical lengths of the resonator be l and θ respectively, and the inductance and capacitance per unit length of the microstrip line be L_0 and C_0 , respectively. With a sinusoidal excitation, the resulting standing waveform is $I = I_0 \sin(\beta x)$. The energy stored in the unloaded system is:

$$U_0 = 0.5 L_0 \int_0^\theta I^2 dx \quad (4.7)$$

It must be noted that θ is a very generalized symbol. It can be expressed in terms of the harmonic frequency of the order mode under investigation. By using this notation, the generality of the theory is preserved.

Expanding equation (4.7):

$$\begin{aligned}
 U_o &= 0.5 L_o \int_0^{\theta} [I_o \sin(\beta x)]^2 \frac{d(\beta x)}{\beta} \\
 &= \frac{1}{8} \frac{Z_o}{\omega} I_o^2 [2\theta - \sin(2\theta)]
 \end{aligned} \tag{4.8}$$

Where $Z_o = (L_o/C_o)^{0.5}$ is the characteristic impedance and $\beta = \omega(L_o C_o)^{0.5}$ is the propagation constant.

When a DUT is inserted at the center of the resonator, the total energy stored in the loaded system is [41]:

$$\begin{aligned}
 U_{sys} &= 0.5 L_o \left\{ \int_0^{\theta/2} I^2 dx + \int_{\theta/2}^{\theta} I^2 dx \right\} + U_{DUT} \\
 &= \frac{1}{8} \frac{Z_o}{\omega} I_o^2 [2\theta - \sin(2\theta)] + U_{DUT}
 \end{aligned} \tag{4.9}$$

Where U_{DUT} is the energy stored in the DUT.

It must be noted that equation (4.9) indicates that the total energy stored in the system equals the sum of the energies stored in all elements, confirming equation (4.5).

4.3.2 Energy Stored in a Thin-film Element

In this study, a thin-film MIM capacitor is under investigation. Examining the equivalent circuit of Figure 3.13, it is found that the series resistance and shunt capacitances contribute very small series impedances compared to those contributed by the series inductance and capacitance. Therefore, it can be safely assumed that almost all the energy is stored in the series inductance and series capacitance. Let C_c and L_c be the net series capacitance and inductance introduced by the MIM structure, the energy stored by this DUT is:

$$U_{DUT} = U_{MIM} = \frac{1}{2} \frac{I_o^2 [\sin(\theta/2)]^2}{\omega^2 C_c} + \frac{1}{2} I_o^2 L_c [\sin(\theta/2)]^2 \tag{4.10}$$

The first term of equation (4.10) represents the energy stored in the capacitive element of the MIM structure. The term becomes less important as the MIM structure becomes more inductive at higher frequencies. The procedure in which L_c and C_c are calculated uses the fact that L_c and C_c are connected in series for a MIM structure. That is:

$$X_{MIM} = \omega L_c - \frac{1}{\omega C_c} \quad (4.11)$$

where X_{MIM} is the reactance of the MIM structure.

Rewriting (4.11):

$$\omega X_{MIM} = \omega^2 L_c - \frac{1}{C_c} \quad (4.12)$$

This is an equation of a straight line when ωX_{MIM} is plotted versus ω^2 ($\omega \neq 0$). With a knowledge of two values of X_{MIM} calculated from the RFS technique, which will be described next, the values of L_c and C_c can be determined.

4.3.3 Resonance Condition of a Loaded Resonator

Figure 4.4 A linear resonator centrally loaded by a zero length DUT

Figure 4.4 depicts a linear resonator loaded by a DUT at the center. The condition for a series resonance is that the impedance of the loaded system becomes zero at a particular frequency. That is:

$$\begin{aligned} Z_1 + Z_L + Z_1 &= 0 \\ \text{or } 2Z_1 + Z_L &= 0 \end{aligned} \quad (4.13)$$

Because Z_1 is the input impedance of an open circuit transformed over a transmission line of length $\theta/2$, it is:

$$Z_1 = -jZ_0 \cot(\theta/2) \quad (4.14)$$

Also, if the DUT has low resistive loss, Z_L is almost purely reactive.

$$Z_L = jX_L \quad (4.15)$$

Using equations (4.13), (4.14) and (4.15), one gets:

$$X_L = 2Z_0 \cot(\theta/2) \quad (4.16)$$

Equation (4.16) signifies the condition at which a series resonance is constituted. This equation is used together with (4.12) to evaluate the values of L_C and C_C .

4.3.4 The Resonant Frequency Shift (RFS) Technique

Figure 4.5 The resonant frequency shift phenomenon

As previously mentioned, a linear resonator, when loaded by a DUT at its center will experience frequency shifts and Q value degradations at all odd-order resonances. This is shown in Figure 4.5, where the solid curve represents the unloaded response whereas the dotted curve represents the shifted response due to central loading. The theories [41] derived thus far

allow the calculation of the DUT's reactance, the individual and total energies stored in the resonant structure and the DUT's Q value, respectively. The only important quantity to be discussed here is θ , the electrical length of the resonator of interest. θ is directly proportional to the order of resonance of the resonant structure. By definition, θ is:

$$\theta \equiv \beta l \equiv \frac{2\pi f}{v} l \quad (4.17)$$

where v is the velocity of wave propagation on the transmission line.

θ can be conveniently expressed in terms of measurable quantities such as unloaded and loaded resonant frequencies, so that it can be directly calculated, as follows.

For the unloaded system with just the resonator:

$$\theta_o = \frac{2\pi f_o}{v} l \quad (4.18)$$

For the loaded system:

$$\theta_{sys} = \frac{2\pi f_{sys}}{v} l \quad (4.19)$$

where f_o and f_{sys} are the unloaded and loaded resonant frequencies, respectively.

For the unloaded end-coupled linear resonator studied here, the length l is related to the propagating wavelength λ_g by:

$$l = n \frac{\lambda_g}{2}$$

where n indicates the order of resonance.

Then:

$$\theta_o = \beta_o l = \frac{2\pi}{\lambda_g} n \frac{\lambda_g}{2} = n\pi \quad (4.20)$$

Using (4.18) and (4.20), one obtains:

$$\frac{2\pi l}{v} = \frac{n\pi}{f_o} \quad (4.21)$$

Putting (4.2) back into (4.19):

$$\theta_{sys} = \frac{f_{sys}}{f_o} n\pi \quad (4.22)$$

With f_{sys} and f_o obtained from the measurements, θ_{sys} is determined by equation (4.22). It is to be noted that θ_o is to be used in equation (4.8) whereas θ_{sys} is to be used in equations (4.10) and (4.16).

4.5 Calculations of the Q Factor and Series Resistive Loss of a DUT

Evidently, a reference (unloaded) resonator can be measured to collect the values of f_o 's and Q_o 's at all resonances. Then a second measurement is performed for the loaded resonator to collect the values of f_{sys} 's and Q_{sys} 's. From the shifts in resonant frequencies and changes in Q values at the odd order resonances, the reactances and Q values of the DUT are computed. The only drawback of this approach is that two passes of measurements are required.

Another approach was suggested which involves only one set of measurements [40]. It involves collecting resonant frequencies and Q values of both odd- and even-order modes for a loaded resonator. Theoretically, the even-order modes are not affected by the central loading. The data obtained for the even-order resonances will thus yield the f_o 's and Q_o 's. The Q_o 's values can be fitted to a linear curve on a log-to-log Q versus frequency plot so that the Q values at the odd-order resonances can be interpolated. The odd-order measurement data are influenced by the loading, hence will yield the f_{sys} 's and Q_{sys} 's. The calculations of the DUT's reactances and Q values can then be performed as follows:

$$\begin{aligned}\theta_o &= n\pi \\ \theta_{sys} &= \frac{f_{sys}}{f_o} n\pi \\ X_L &= 2Z_o \cot(\theta_{sys}/2) \\ \omega X_L &= \omega X_{MIM} = \omega^2 L_c - \frac{1}{C_c} \\ U_{MIM} &= \frac{1}{2} \frac{I_o^2 [\sin(\theta_{sys}/2)]^2}{\omega^2 C_c} + \frac{1}{2} I_o^2 L_c [\sin(\theta_{sys}/2)]^2 \\ U_{sys} &= \frac{1}{8} \frac{Z_o}{\omega} I_o^2 [2\theta_{sys} - \sin(2\theta_{sys})] + U_{MIM} \\ U_o &= \frac{1}{8} \frac{Z_o}{\omega} I_o^2 [2\theta_o - \sin(2\theta_o)] \\ \frac{U_{sys}}{Q_{sys}} &= \frac{U_o}{Q_o} + \frac{U_{MIM}}{Q_{MIM}}\end{aligned}$$

Therefore:

$$Q_{MIM} = \left[\frac{1}{U_{MIM}} \left(\frac{U_{sys}}{Q_{sys}} - \frac{U_o}{Q_o} \right) \right]^{-1} \quad (4.23)$$

And:

$$R_{MIM} = \frac{X_{MIM}}{Q_{MIM}} \quad (4.24)$$

where

n = order of resonance

Z_o = characteristic impedance of the microstrip resonator

R_{MIM} = effective series resistive loss of the MIM capacitor

4.6 Summary

An on-substrate linear resonant structure is described in this chapter to accommodate the requirements of Q factor and low series resistive loss measurements at microwave frequencies. The technique eliminates the use of bonding wires, thus reduces the repeatability problem significantly. Also, the technique employs an indirect calculation, for the Q factor of the DUT, which makes the results less sensitive to the measurement fluctuations. This feature is particularly advantageous especially for frequencies in the upper end of the Ka band and beyond (26 GHz to 40 GHz). The only data directly obtained from the measurements are the resonant frequencies and Q values, at all resonances, of the center-loaded resonator. The measured even-order resonances are not affected by the central loading, and hence represent the unloaded characteristics of the resonator. From the even-order mode measurements, the unloaded Q values (Q_o 's) at the odd-order modes are interpolated. The measured odd-order resonances, on the other hand are affected by the loading, allowing the loaded Q values (Q_{sys} 's) of the resonator to be determined. The DUT's Q values and series resistive losses are then evaluated from the differences in Q values and resonant frequencies of the odd-order modes due to the insertion of the DUT.

CHAPTER 5: RESULTS AND DISCUSSION

5.1 Introduction

In this chapter, the results of the complete characterization of thin-film MIM capacitors are presented. These include DC measurements, RF measurements, capacitor modellings and Q-factor measurements as described in the previous chapters.

5.2 DC Measurements of Thin-film MIM Capacitors

Two batches of MIM capacitors were examined using the HP 4280 CV-meter, at 1 MHz. The first batch was processed using silicon dioxide, SiO₂, and the second batch using silicon nitride, Si₃N₄, as dielectric materials. The values of the SiO₂ MIM capacitors ranged from 0.1 pF to 50 pF whereas the Si₃N₄ MIM capacitors were from 0.8 pF to 20 pF. The results, of the DC evaluation, are as shown in Table 5.1 and 5.2, assuming the relative dielectric constants of SiO₂ and Si₃N₄ are 4.8 and 7, respectively.

Table 5.1
Capacitances of SiO₂ MIM Capacitors at 1 MHz

Designed area (width x length) [mil x mil]	Designed capacitance [pF]	Measured area (width x length) [mil x mil]	Measured capacitance [pF]
27 x 27	50	27.1 x 27.1	67.775
12 x 12	10	12.17 x 12.17	13.715
10.8 x 10.8	8	10.87 x 10.87	11.290
9.3 x 9.3	6	9.63 x 9.63	8.350
7.6 x 7.6	4	7.75 x 7.75	5.690
5.4 x 5.4	2	5.49 x 5.49	3.300
3.8 x 3.8	1	3.75 x 3.75	1.855
3.4 x 3.4	0.8	3.48 x 3.48	1.475
2.95 x 2.95	0.6	2.81 x 2.81	1.290
2.4 x 2.4	0.4	2.34 x 2.34	0.97
1.7 x 1.7	0.2	1.55 x 1.55	0.69

Table 5.2
Capacitances of Si₃N₄ MIM Capacitors at 1 MHz

Designed area (width x length) [mil x mil]	Designed capacitance [pF]	Measured area (width x length) [mil x mil]	Measured capacitance [pF]
250 x 250	20.375	250 x 250	19.13
109 x 109	3.873	109 x 109	3.63
95 x 95	2.942	95 x 95	2.75
75 x 80	1.956	75 x 80	1.83
55 x 55	0.986	55 x 55	0.92
50 x 50	0.815	50 x 50	0.77

The measurements of SiO₂ capacitors, fabricated at Carleton University, indicated that the actual capacitances were much higher than those originally designed. In this batch of capacitors, the thickness of the SiO₂ layer was approximately 4300Å which was slightly more than the desired value at 4000Å. This difference did not, however, account for the high capacitances measured. The capacitances should have been lower with greater dielectric layer thickness. In order to identify the source(s) of error(s), the HP4280 CF-meter accuracy and the assumed value of SiO₂ relative dielectric constant were verified. Results of this study indicated that performance of the CV-meter was satisfactory and the relative dielectric constant of SiO₂ was approximately about 4.8. Thus, it had been concluded that the surface roughness of the alumina substrates, at 3 µinch, was the cause of the variation in capacitance values. Calculations performed using reference [8] revealed that a 3 µinch (762Å) surface finish substrate would contain randomly situated craters as large as 1.2 µm in radius. Due to such poor surface finish, the effective capacitor areas were much larger than the top electrode areas measured under a microscope. The smaller the capacitor dimensions, the larger the differences would be between designed and measured capacitance values, as indicated in Table 5.1.

The second batch of capacitors were fabricated on a very smooth surface finish, at 1 µinch, to correct the above problem. The dielectric material was changed to Si₃N₄ because the fabrication facility was switched to that of CRC. The change in fabrication facility was due to the fact that the author joined CRC at the same time the second batch was started. It must be noted that there

was no apparent preference to Si_3N_4 over SiO_2 . Si_3N_4 was used due to its higher dielectric constant and availability. As shown in Table 5.2, the results were much improved. The thickness of the deposited Si_3N_4 layer was measured to be about $2000\text{\AA}$. The MIM capacitors were laid out on a wrap-around-ground configuration, as shown in Figure 5.1, so that they could later be probed by the Cascade-Microtech wafer prober.

Figure 5.1. Typical MIM capacitor cell for microwave wafer probing.

This test cell is very convenient for microwave probing but the length of the cell must be kept short so that excessive inductances, introduced by ground patches, remain insignificant at high frequencies, especially above 20 GHz.

5.3 RF Measurements of Thin-film MIM Capacitors

Si_3N_4 MIM capacitors, fabricated in subsequent batches, were microwave probed on the Hewlett Parkard (HP) 8510 automatic network analyzer (ANA) to determine their S-parameters, in the frequency range from 0.5 GHz to 26.5 GHz. TRL calibration was performed to de-embed the systematic errors of the ANA.

5.3.1 TRL Calibration

Calibration standards utilized in the TRL technique were those which employed also the wrap-around-ground configuration to correctly simulate the operating environment of MIM capacitors during measurements. The standards include a through line, an open and a delay line which possessed the same length as that of a MIM capacitor cell. They are as shown in Figure 5.2.

Figure 5.2. TRL calibration standards
(a) Through; (b) Reflect; (c) Line (delay)

The TRL calibration was verified by measuring two transmission lines, in the wrap-around-ground configuration, with one longer and the other shorter than the delay standard. Both test lines were, however, longer than the through standard. The results obtained were better than 35 dB return loss and ± 0.1 dB insertion loss for both lines, in the 0.5-26.5 GHz frequency range. These figures were satisfactory for S-parameter measurements to be reliably performed.

It must be noted that the TRL calibration described here was performed in the wrap-around-ground environment, thus, it was applicable only to that particular group of measurement cells. For measurement cells which employed via-through grounds, a different set of standards must be employed to appropriately de-embed the ANA.

5.3.2 S-parameter Measurements

Si₃N₄ MIM capacitors, ranging from 0.1 pF to 20 pF, were measured on the HP 8510 ANA. Only measurements of the 0.5 pF and the 3.9 pF, on

alumina, and the 20 pF, on quartz, capacitors are shown here for illustrative purposes. Figures 5.6, 5.7 and 5.8, in section 5.4.1, display the measured responses of S_{11} and S_{21} for the three capacitors, respectively. It is to be noted that Figures 5.6, 5.7 and 5.8 also include the responses of the proposed physical model and Libra's thin-film capacitor (TFC) model to verify the validity of the proposed model.

The responses all indicate that the MIM capacitors exhibit series resonances in the frequency ranges of interest. The series resonance is clearly demonstrated by a sharp fall in the $|S_{11}|$ response. Beyond the series resonant frequency, the MIM capacitor becomes inductive and higher insertion loss is experienced. The series resonant frequency is inversely proportional to the capacitance value, as expected. Therefore, in general, when MIM capacitors are used for matching purposes, the operating frequency must be kept below that of the lowest series resonant frequency. A violation of this rule may prove to be disastrous as the MIM capacitors no longer behave capacitively. However, in some special cases, if correctly modelled, these MIM capacitors can be employed in microwave applications in which inductors are replaced by inductive capacitors. Extreme caution must be taken in such situations because of the complexity and eccentricity of the applications.

The first parallel resonance of the MIM capacitors under test was not detected in the 0.5 - 26.5 GHz frequency range. Thus, the model for MIM capacitors to be developed in the immediately following section could not be tested for its validity in the vicinity of the first parallel resonance. Even so, a model using the lumped physical modelling technique, was developed to study the responses of MIM capacitors in the 0.5-26.5 GHz frequency range.

5.4 Modelling of MIM Capacitors

5.4.1 Parameter Calculations

As proposed in Section 3.4.3, the physical lumped modelling was applied to a group of Si_3N_4 MIM capacitors ranging from 0.1 pF to 20 pF, with dielectric thickness of 2000 Å. Only the results of three MIM capacitors, 0.5 pF, 3.9 pF and 20 pF, are presented here. The dimensions and geometries of these are given in Figures 5.3, 5.4 and 5.5, respectively. The contributions of the elementary discontinuities, which form the capacitor, are derived according to capacitor geometry and are listed in Tables 5.3, 5.4 and 5.5, respectively.

These discontinuities include all stray coupling capacitances, series inductive and resistive losses and coupling capacitance. The subroutines, written in Mathematica™, which was programmed to perform equivalent circuit derivations for all discontinuities, are given in Appendix IV.

Figure 5.3 Geometry of a 0.5 pF MIM capacitor

Table 5.3
Contributions of a 0.5 pF MIM Capacitor's Discontinuities

Element	Dimensions	Contribution
Parallel plate structure	40 μm x 38 μm	series 0.471 pF
Input microstrip step	$W_{\text{big}}=74 \mu\text{m}$ $W_{\text{small}}=48 \mu\text{m}$	shunt 0.0002 pF series 0.0116 nH
Output microstrip step	$W_{\text{big}}=74 \mu\text{m}$ $W_{\text{small}}=10 \mu\text{m}$	shunt 0.0019 pF series 0.0642 nH
Microstrip gap	gap=19 μm width=48 μm	shunt 0.0005 pF coupling 0.0547 pF shunt 0.0005 pF
Airbridge	19 μm x 10 μm	series 0.0083 nH series 0.0250 Ω
Electrode	48 μm x 38 μm	series 0.0149 nH series 0.0178 Ω

Figure 5.4 Geometry of a 3.9 pF MIM capacitor

Table 5.4
Contributions of a 3.9 pF MIM Capacitor's Discontinuities

Element	Dimensions	Contribution
Parallel plate structure	109 μm x 109 μm	series 3.6818 pF
Input microstrip step	$W_{\text{big}}=118 \mu\text{m}$ $W_{\text{small}}=74 \mu\text{m}$	shunt 0.0004 pF series 0.0122 nH
Output microstrip step	$W_{\text{big}}=74 \mu\text{m}$ $W_{\text{small}}=10 \mu\text{m}$	shunt 0.0019 pF series 0.0642 nH
Microstrip gap	gap=16 μm width=74 μm	shunt 0.0006 pF coupling 0.0763 pF shunt 0.0006 pF
Airbridge	16 μm x 10 μm	series 0.0065 nH series 0.0211 Ω
Electrode	118 μm x 109 μm	series 0.0287 nH series 0.0156 Ω

Figure 5.5 Geometry of a 20 pF MIM capacitor

Table 5.5
Contributions of a 20 pF MIM Capacitor's Discontinuities

Element	Dimensions	Contribution
Parallel plate structure	250 μm x 250 μm	series 19.3681 pF
Input microstrip step	$W_{\text{big}}=262 \mu\text{m}$ $W_{\text{small}}=74 \mu\text{m}$	shunt 0.0038 pF series 0.0316 nH
Output microstrip step	$W_{\text{big}}=74 \mu\text{m}$ $W_{\text{small}}=20 \mu\text{m}$	shunt 0.0011 pF series 0.0331 nH
Microstrip gap	gap=40 μm width=74 μm	shunt 0.0010 pF coupling 0.0667 pF shunt 0.0010 pF
Airbridge	40 μm x 20 μm	series 0.0168 nH series 0.0276 Ω
Electrode	262 μm x 250 μm	series 0.0536 nH series 0.0151 Ω

The elementary discontinuities, listed in Table 5.6, 5.7 and 5.8, can be combined in the manner described in Section 3.4.3 to construct the equivalent circuits of the three MIM capacitors as proposed by the physical lumped modelling technique. The resulting model is illustrated in Figure 3.13 and the model parameters are as summarized in Table 5.6 below.

Table 5.6
Parameters of MIM Capacitor Models (with reference to Figure 3.13)

MIM capacitor [pF]	L [nH]	C ₁ [pF]	R [Ω]	C ₂ [pF]	C ₃ [pF]	C ₄ [pF]
0.5	0.113781	0.471	0.06074	0.05465	0.00075	0.00243
3.9	0.140132	3.682	0.05220	0.07632	0.00094	0.00244
20	0.188655	19.368	0.05789	0.06667	0.00485	0.00218

The performances of these models are compared to S-parameter measurements and the thin-film capacitor (TFC) model available in Libra™. The comparisons are shown in Figures 5.6, 5.7 and 5.8.

Figure 5.6 S₁₁ and S₂₁ responses of the measurement, the physical model and Libra's TFC model of a 0.5 pF MIM capacitor on alumina
(see legend next page)

Figure 5.7 S₁₁ and S₂₁ responses of the measurement, the physical model and Libra's TFC model of a 3.9 pF MIM capacitor on alumina (see legend below)

Figure 5.8 S₁₁ and S₂₁ responses of the measurement, the physical model and Libra's TFC model of a 20 pF MIM capacitor on quartz

Legend:

○ S ₁₁ measure	□ S ₂₁ measure
◆ S ₁₁ model	▲ S ₂₁ model
— S ₁₁ TFC	● S ₂₁ TFC

5.4.2 Discussion

Overall, the models proposed agreed well with measurements in the 0.5 GHz to 26.5 GHz frequency range. The models indicated the same series resonances as those of the measurements. Both magnitude and phase responses proved the validity of the models. The errors in phase responses of the physical models were more profound than that of the magnitude responses, especially near the series resonant frequency. This was as expected because the physical models were lumped, and thus, could not fully describe all distributed effects which occur in the MIM capacitors. It was also found that the physical model performed much better for the cases in which the parallel plate capacitances were more dominant than the parasitic elements. A comparison between Figure 5.10 or 5.11 to Figure 5.9 would verify this observation.

Moreover, comparisons were also made among the proposed models, Libra™ TFC models and the measurements, for three capacitors. It appeared that the Libra™ TFC models did not predict the series resonances observed in the $|S_{11}|$ responses as indicated by the measurements and the proposed models. The Libra™ TFC model may thus be disastrous for applications in which MIM capacitors are employed beyond their misrepresented first series resonant frequencies. Due to this misrepresentation, the overall performance of a microwave integrated circuit may be incorrectly simulated.

This modelling exercise was also performed to characterize MIM capacitors up to their first parallel resonances, at which the transmission responses of these capacitors would diminish. However, due to the low-loss characteristics of these MIM capacitors, measurements performed from 0.5 GHz to 26.5 GHz did not detect any parallel resonance. Also, subsequent simulations of the physical models using Libra did not predict any parallel resonances in this frequency range either. Therefore, at this stage, no conclusion can be made about the validity of the proposed model in the vicinity of the first parallel resonance. Higher frequency measurements are to be performed or more lossy MIM capacitors are to be examined in order to study the effectiveness of the physical lumped modelling technique.

The algorithm, from which model parameters are derived, can be easily programmed to a user-defined library for public access because all formulae involved are closed-form expressions. Since no optimization against measurements was necessary in the development of the lumped

models, capacitor geometry synthesis is allowed in such a user-defined MIM model. Currently, a user defined element representing a MIM capacitor is being adapted to Libra™ to facilitate general simulation and optimization requirements.

5.5 Q-factor Measurements of MIM Capacitors

The RFS procedure summarized in Section 4.5 is applied here to determine Q-values of MIM capacitors. In order to more clearly illustrate the loading effects of a series connected MIM capacitor on a linear resonator, small and moderate capacitance values were chosen. At a particular frequency, smaller capacitors exhibit higher impedances, hence, more profoundly perturb the resonator. In this study, seven nearly square MIM capacitors of values 0.6 pF, 1.5 pF, 3 pF, 5 pF, 11 pF, 17 pF and 23 pF, were analyzed. Only the sample calculations of the 0.6 pF and 5 pF capacitors are presented here for illustrative purposes. The Mathematica™ program written to perform these calculations is given in Appendix V for future reference.

Due to fabrication restrictions, only 1x1 in² substrates were used in the MHMIC process. The longest end-coupled linear resonator which could be fabricated on these substrates corresponds to that of a 2.85 GHz resonator. It must be noted that the lower the fundamental resonant frequency the more harmonic data points one could obtain from the RFS technique. In the frequency range from 1 GHz to 40 GHz, a 2.85 GHz resonator would provide about 14 harmonic data points, which allowed 7 data points for each of the loaded and unloaded cases.

Figure 5.9 Layouts of RFS elements with different dimensions

- a) small capacitor
- b) large capacitor

The total length of the resonant structure, which includes the resonator, the MIM capacitor and two feedlines, was kept constant at 875 mil to be used with the Intercontinental test fixture available. The lengths of the feedlines were adjusted in each case to accommodate the changes in capacitor size. Two resonant structures are shown in Figure 5.9 to illustrate typical arrangement of a Q-factor measurement mask.

5.5.1 Transmission Measurements

Typical transmission responses of an ended-coupled linear resonator, loaded with a 0.6 pF and a 5 pF MIM capacitor at the middle, are as shown in Figures 5.10 a) and b), respectively. As expected, the 0.6 pF MIM capacitor perturbs the resonator more than the 5 pF one and shifts the frequencies of the odd-order harmonics more noticeably. As described in Chapter 4, the Q-values and frequencies of the even-order harmonic modes represent the unloaded characteristics of the resonator whereas those of the odd-order harmonic modes represent the loaded behaviours. It must be noted that there are two types of loading effect which occur within the linear resonant structure in the RFS technique. The primary loading effect is due to perturbation of the DUT. The secondary loading effect is due to the gap couplings at both sides of the resonant structure. In this thesis, unless stated otherwise, the terms loaded and unloaded are used only to signify the primary loading effect.

All Q-measurements were performed using the Wiltron 360 ANA and the Intercontinental universal test fixture. Initially, a broadband TRL calibration, from 1 GHz to 40 GHz, was carried out to obtain a transmission response as that shown in Figure 5.10 a) or b). Sub-band TRL calibrations, with 2 GHz bandwidth, were then performed to study the transmission response at each harmonic frequency. Using 501 data points, a resolution of 4 MHz was achieved. This was sufficient for microstrip Q-value determination. For a resonator fundamental frequency of 2.85 GHz, 14 sub-band calibrations were needed to cover the 14 harmonics displayed in the 1 GHz to 40 GHz frequency range. These calibrations strongly influenced the Q-measurements therefore they were extensively verified.

(a)

(b)

Figure 5.10 Measured transmission responses of resonant structures
a) loaded by a 0.6 pF MIM capacitor
b) loaded by a 5 pF MIM capacitor

5.5.2 Unloaded characteristics of an end-coupled linear resonator

The RFS measurements were carried out one resonant structure at a time to ensure the consistency of connector contacts to microstrip feedlines. The resonant structure was held by the Intercontinental test fixture and subjected to 14 harmonic sub-band measurements. For each measurement, the order of resonance, the resonant frequency, the magnitude of the transmission coefficient and the 3-dB bandwidth were collected. The Q-factors of the resonant structure were then determined as a function of harmonic order, or in other words, as a function of frequency. Tables 5.7 a) and b) below show the results obtained from the RFS measurements for all harmonics, but only even-order harmonics are examined in this section to study the unloaded behaviour of the resonator.

Table 5.7
a) Harmonic Responses of 0.6 pF MIM Capacitor

n	Δf [GHz]	m [dB]	f_0 [GHz]	β	Q_0
1	4.108-4.060	-33.723	4.084	0.010562	86.8728
2	5.669-5.615	-26.309	5.640	0.0254127	109.753
3	9.095-9.021	-22.148	9.060	0.0423528	132.803
4	11.262-11.186	-18.919	11.224	0.0638587	166.546
5	14.388-14.292	-17.008	14.340	0.082156	173.919
6	16.772-16.672	-14.862	16.720	0.110259	204.071
7	19.718-19.593	-15.231	19.656	0.104713	190.180
8	22.199-22.072	-13.008	22.136	0.144053	224.516
9	25.011-24.852	-12.812	24.932	0.148315	203.318
10	27.548-27.394	-12.926	27.472	0.145818	230.414
11	30.317-30.108	-12.932	30.184	0.145688	186.502
12	32.964-32.709	-12.143	32.788	0.164087	170.777
13	35.487-35.212	-10.406	35.356	0.216113	184.138
14	37.968-37.670	-10.800	37.876	0.202645	178.613

b) Harmonic Responses of a 5 pF MIM Capacitor

n	Δf [GHz]	m [dB]	f_0 [GHz]	β	Q_0
1	3.012-2.944	-38.186	2.976	0.00623814	44.3107
2	5.575-5.495	-26.808	5.533	0.023926	72.4760
3	8.375-8.282	-21.181	8.326	0.0478174	98.0888
4	11.0847-10.9848	-17.243	11.036	0.0796139	128.060
5	13.789-13.674	-14.995	13.732	0.108221	145.254
6	16.521-16.395	-12.910	16.458	0.146165	168.803
7	19.1498-19.0006	-12.646	19.073	0.152047	166.709
8	21.871-21.7154	-10.574	21.796	0.210233	198.975
9	24.4223-24.2348	-10.152	24.330	0.225418	188.260
10	27.1574-26.9663	-9.981	27.060	0.231979	207.298
11	29.652-29.435	-9.502	29.532	0.251753	204.616
12	32.4423-32.189	-10.053	32.280	0.229186	185.852
13	34.7965-34.4942	-8.828	34.688	0.283588	179.829
14	37.440-37.1653	-8.629	37.336	0.294025	215.841

where

n is the harmonic order

Δf is the 3dB-bandwidth of the transmission response at resonance

m is the magnitude of the transmission response at resonance

f_0 is the resonant frequency

β is the microstrip-gap coupling coefficient

and $\beta = \{10^{(m/20)}\} / \{2(1 - 10^{(m/20)})\}$ for a low-loss coupling structure

Q_0 is the unloaded Q-factor

In Tables 5.7 a) and b), the gap coupling coefficients, β , at both input and output ports were identical due to the symmetry of the resonant structure. The Q-factors, Q_0 , of the resonant structure were corrected from the gap coupling, which was the secondary loading effect, by using Equation 5.1 [45].

$$Q_0 = \frac{f_0}{\Delta f} (1 + 2\beta) \quad (5.1)$$

where: Q_0 is the unloaded Q-value

f_0 is the unloaded resonant frequency (even-order harmonics)

As suggested in Chapter 4, the even-order Q-values and resonant frequencies were fitted to a linear log-to-log function. The data spread and fitted functions, of a 0.6 pF and a 5 pF MIM capacitors, are as shown in Figure 5.11.

Figure 5.11 Measured unloaded Q-responses and fitted functions of
a) resonator perturbed by a 0.6 pF MIM capacitor
b) resonator perturbed by a 5 pF MIM capacitor

It must be noted that the slopes of the fitted linear functions, shown in Figure 5.11, profoundly influenced the results of subsequent calculations such as MIM Q-value calculations. These functions were fitted more closely to the low frequency data points than the high frequency ones. Due to excessive radiation at high frequencies, especially beyond 30 GHz, high even-order harmonic data points did not truly reflect the nature of the unloaded resonator, therefore, could not be fully utilized. In other words, the lower even-order harmonic data points must have higher weights in formulating the above fitted functions.

Figures 5.11 a) and b) show the unloaded Q responses of two resonant structures perturbed by a 0.6 pF and a 5 pF MIM capacitors, respectively. As displayed, the log-to-log linear functions are fitted mainly to the first few even-order harmonic points. The expressions are as shown on the figures. With these expressions, the odd-order unloaded Q-values and resonant frequencies can be interpolated for subsequent usage.

5.5.3 Reactance of a Perturbing MIM Capacitor

From the frequency shifts which occurred at the odd-order resonances, the reactance of the perturbing DUT, which was a MIM capacitor in this case, could be calculated. This calculation required the knowledge of both the unloaded and loaded resonant frequencies of a particular odd-order resonance. The unloaded resonant frequencies were interpolated from the even-order measurements as described in section 5.5.2. The loaded resonant frequencies were obtained directly from the RFS measurements. Tables 5.8 a) and b) show the unloaded (f_0) and loaded (f_{sys}) resonant frequencies, the electrical length (θ_{sys}) and the reactance (X_L or X_{MIM}) of the 0.6 pF and the 5 pF MIM capacitors, respectively, at all odd-order resonances. θ_{sys} and X_L (or X_{MIM}) were calculated using Equations (4.22) and (4.16), respectively.

It is to be noted that the frequency at which the reactance value changes from negative to positive is the MIM capacitor's first self series resonant frequency. At this frequency, the capacitor does not behave capacitively but rather inductively.

Table 5.8

a) Electrical Lengths and Reactances for a 0.6 pF MIM Capacitor

n	f_0 [GHz]	f_{sys} [GHz]	θ_{sys} [radian]	X_L [Ω]
1	2.820	4.084	4.54974	-84.9275
3	8.432	9.060	10.1244	-36.4812
5	13.972	14.340	16.1179	-20.7903
7	19.428	19.656	22.2451	-12.7688
9	24.804	24.932	28.4169	-7.14196
11	30.130	30.184	34.6177	-3.00882
13	35.332	35.356	40.8674	-1.33693

b) Electrical Lengths and Reactances for a 5 pF MIM capacitor

n	f_0 [GHz]	f_{sys} [GHz]	θ_{sys} [radian]	X_L [Ω]
1	2.76665	2.976	3.37931	-11.9424
3	8.28465	8.326	9.47173	-2.34812
5	13.7470	13.732	15.6909	0.851662
7	19.1270	19.073	21.9298	3.06687
9	24.4280	24.330	28.1631	5.56981
11	29.6700	29.532	34.4008	7.85118
13	34.8080	34.688	40.7044	6.82370

5.5.4 Energies Stored in a Resonant Structure

With the determination of θ_{sys} , the energies stored in the linear resonator and the perturbing MIM capacitor can be computed using Equation (4.8) and (4.10), respectively. Since the current, I_0 , is common in every term, it can be normalized from both equations. Equation (4.8) is simply a function of one unknown θ_{sys} . Equation (4.10), on the hand, involves also two other unknowns, L_c and C_c . In order to determine the energy stored, per unit current, in the perturbing MIM capacitor, the values of these two unknowns must be computed using the method described in section 4.3.2. The

reactances X_L (or X_{MIM}) and loaded resonant frequencies were fitted to Equation (4.11) and the values of L_C and C_C were calculated.

Figure 5.12 Measured reactances and fitted functions of
a) 0.6 pF MIM capacitor
b) 5 pF MIM capacitor

Figure 5.12 shows the fitted curve and function of X_L (or X_{MIM}) versus frequency against the measured data. The fitted function is of the form of that shown in Equation (5.2).

$$X = af - \frac{b}{f} \quad (5.2)$$

where f is frequency in GHz
 a and b are fitted coefficients

The net inductance and net capacitance deduced from Equation (5.2) are shown in Equation (5.3) and (5.4).

$$L_c = \frac{a}{2\pi} 10^{-9} \quad (5.3)$$

$$C_c = \frac{1}{2\pi b} 10^{-9} \quad (5.4)$$

For the 0.6 pF MIM capacitor under investigation here, the net inductance, L_c , and net capacitance, C_c , calculated from the fitted equation in Figure 5.12 a) were 0.042 nH and 0.453 pF, respectively. Similarly, for the 5 pF MIM capacitor, the values were 0.043 nH and 4.210 pF. These values were very close to the values of L and C_1 derived from the physical modelling technique, which implied the dominance of these two elements. The energy stored in the MIM capacitor of interest must be the sum of the energies stored in the net inductance and the net capacitance as suggested by Equation (4.10).

Table 5.9

a) Stored Energies (Per unit Squared Current) of a Resonator Perturbed by a 0.6 pF MIM Capacitor

n	U_{MIM} $\times 10^{-9} [J/A^2]$	U_o $\times 10^{-9} [J/A^2]$	U_{sys} $\times 10^{-9} [J/A^2]$	Q_o (at f_o)	Q_{sys} (at f_{sys})
1	0.492987	2.13847	2.63146	77.3649	86.8728
3	0.159555	2.11498	2.27454	138.681	132.803
5	0.0752227	2.18538	2.26060	181.505	173.919
7	0.0459292	2.22688	2.27280	216.361	190.180
9	0.0328186	2.25629	2.28911	246.441	203.318
11	0.0258180	2.27771	2.30352	273.355	186.502
13	0.0216771	2.29805	2.31973	297.567	184.138

b) Stored Energies (Per unit Squared Current) of a Resonator Perturbed by a 5 pF MIM Capacitor

n	U_{MIM} $\times 10^{-9} [\text{J}/\text{A}^2]$	U_0 $\times 10^{-9} [\text{J}/\text{A}^2]$	U_{sys} $\times 10^{-9} [\text{J}/\text{A}^2]$	Q_0 (at f_0)	Q_{sys} (at f_{sys})
1	0.177951	2.10605	2.28401	42.5189	44.3107
3	0.0323419	2.25200	2.28434	100.138	98.0888
5	0.0186378	2.27571	2.29434	148.722	145.254
7	0.0147835	2.29380	2.30859	192.488	166.709
9	0.0131629	2.31189	2.32505	233.014	188.260
11	0.0123115	2.32781	2.34012	271.221	204.616
13	0.0118576	2.34221	2.35407	307.253	179.829

5.5.5 Q-factor Calculations

The results obtained thus far are for the two 0.6 pF and 5 pF capacitors. Similar results for 1.5 pF, 3 pF, 11 pF, 17 pF and 23 pF MIM capacitors were also obtained. Using these results and Equations (4.23) and (4.24), the Q-factors of all seven MIM capacitors were computed as functions of frequency.

Using the physical modelling technique described in Chapter 3, the models of these seven capacitors were developed. From the models, the Q-factor responses, as functions of frequency, were established and compared to the results obtained from the RFS technique above. By definition, Q-factor of an element at a particular frequency is the ratio of the total energy stored in the element to its power dissipated. This definition is used to confirm the agreement (or disagreement) between the devised modelling technique and the RFS technique, outside of the immediate neighbourhood of the element's self series resonance. The series resistance component of the model was assumed to be proportional to the skin depth [46] which was a function of $\sqrt{f}$. This assumes the loss is due entirely to conductor loss and ignores the dielectric loss. Figure 5.13 displays the differences between the physical modelling (model) and RFS (measure) techniques for the 0.6 pF and 5 pF capacitors. Similarly, Figure 5.14 displays the differences for the 3 pF and 23 pF capacitors; Figure 5.15 displays the differences for the 1.5 pF and the 17 pF capacitors; and Figure 5.16 displays the differences for the 11 pF capacitor.

Figure 5.13 Q-factor determination: RSF measurement technique vs. physical modelling for a 0.6 pF and a 5 pF MIM capacitor

Figure 5.14 Q-factor determination: RSF measurement technique vs. physical modelling for a 3 pF and a 23 pF MIM capacitor

Figure 5.15 Q-factor determination: RSF measurement technique vs. physical modelling for a 1.5 pF and a 17 pF MIM capacitor

Figure 5.16 Q-factor determination: RSF measurement technique vs. physical modelling for a 11 pF MIM capacitor

5.5.6 Discussion

The RFS technique employed in this study was proven to be effective in broadband characterization of thin-film MIM capacitors. The results obtained agreed reasonably well with published literature [40], [41], [47]. Comparisons between the Q-values measured using this method and those derived from the physical modelling technique showed that the latter still needed further improvement, especially in calculating the series resistance element in the model. As indicated in Figure 5.13, 5.14, 5.15 and 5.16, the Q-responses all showed that the series resistance element of the physical model varied not simply as a function of $\sqrt{f}$ but in a much more complicated manner. Pettenpaul et al [36] had suggested a more elaborated closed-form formula which could be used to provide a better description of conductor resistance variation as a function of frequency. Also, the contribution of dielectric loss must be taken in account in calculating the value of the total series resistance element.

In the RFS technique, the unloaded odd-order parameters, such as frequency and Q-value, were interpolated from the measured even-order parameters. This interpolation was necessary because both the unloaded and loaded responses of the resonant structure were obtained from the same set of RFS measurements. The even-order measurements yielded the unloaded characteristics whereas the odd-order measurements yielded the loaded characteristics. Such convenience achieved during measurements gave rise to a new uncertainty in the subsequent calculations. In other words, in the RFS technique the repeatability of the measurements was ensured at the price of complicating the analysis. However, the analysis could be easily modified to accommodate the trade-off but the measurements could not. The author had investigated the alternative of measuring a reference resonator separately to obtain the unloaded characteristics of the resonant structure and found that this approach did not yield satisfactory results. One of the obvious problems with the latter approach was the fact that the physical lengths of the reference and the perturbed resonant structures were entirely different. It could be safely stated that the former measurement approach were preferred over the latter even though care must be exercised in interpolating the unloaded responses of the odd-order modes from the even-order information.

It must be noted that the RFS technique required accurate calibration for every harmonic sub-band measurement. In this study, the TRL calibration

technique was employed but, if possible, in the future the broad band TAN technique could be used instead to improve measurement accuracy. Also, measurement resolution better than 4 MHz would definitely enhance the accuracy of the RFS technique.

5.6 Summary

In this chapter, the results of thin-film MIM capacitor's DC, RF, Q-value measurement and modeling techniques were presented. Wherever possible, these measured results were compared to those theoretically derived or supplied by other techniques. Difficulties encountered during measurements or calculations were identified and discussed. Recommendations to improve these techniques were also cited.

The DC measurements for both silicon dioxide and silicon nitride MIM capacitors were performed at 1 MHz using the HP 4280 CV-meter. From the first batch, the silicon dioxide MIM capacitors, fabricated on 3 μ inch surface finish alumina substrates, indicated much higher capacitances than the designed values. Error as large as 245% was found. In order to correct this problem, smoother substrates were employed in the fabrication process. The second batch which produced silicon nitride MIM capacitors were fabricated on 1 μ inch surface roughness substrate and yielded much better agreements with the designed capacitances. Only a 6.2% discrepancy was found, for the worst case.

The S-parameter measurements were performed only for the silicon nitride MIM capacitors, which ranged from 0.1 pF to 20 pF. The HP 8510 ANA and the Cascade-Microtech wafer prober were used to carry out the measurements in the frequency range from 0.5 GHz to 26.5 GHz. The TRL calibration technique was employed to correct the ANA's systematic errors. Observed responses indicated that all MIM capacitors under test exhibited self series resonances within this frequency range. The self resonant frequency was inversely proportional to the capacitance of the parallel plate structure. No parallel resonance, however, was observed due to the low inductive loss nature of these silicon nitride MIM capacitors. Only the results of the 0.5 pF, 3.9 pF and the 20 pF MIM capacitors were presented.

The electrical lumped models of thin-film MIM capacitors, which ranged from 0.1 pF to 20 pF, were developed from their geometries to verify the physical modelling technique. Detailed derivations of these physical

models of the 0.5 pF, the 3.9 pF and the 20 pF capacitors were given in this section to illustrate the technique. Comparisons between the model's responses and the S-parameter measurements, which were provided in the previous section, indicated that the proposed physical modelling technique behaved satisfactorily. Comparisons between the physical model and Libra's TFC model indicated that the TFC model could not accurately characterize thin-film MIM capacitor without prior knowledge of the capacitor's Q-response. This modelling technique is currently being applied to various other thin-film passive elements.

Lastly, the Q-values of the 0.6 pF, 1.5 pF, 3 pF, 5 pF, 11 pF, 17 pF and 23 pF capacitors were measured using the RFS technique, in the frequency range from 1 GHz to 40 GHz. Using a 2.85 GHz end-coupled linear resonator, perturbed in the middle by a MIM capacitor, the RFS technique allowed 14 resonances to be observed, in which 7 odd-order resonances yielded the loaded responses and the 7 even-order resonances yielded the unloaded responses. The Q-values of the MIM capacitors, determined from the RFS technique, seemed to decrease from 120 near 3 GHz to 1 near 37 GHz, in a manner which was of the form $f^{(a-b)}$ [36], and were inversely proportional to the capacitances. The results also indicated that the series resistances of these MIM capacitors were mainly due to conductor losses, for low capacitance values, and became more dominated by dielectric losses, for high capacitance values.

CHAPTER 6: CONCLUSIONS

6.1 Summary and Conclusions:

In this thesis, three topics directly related to the development of a thin-film fabrication process were studied. They are as follows:

1) A design procedure for thin-film MIM capacitors was established using existing facilities at Carleton University. It includes:

a) Simulation Tools:

- i) Frequency domain simulation;
 - + Linear: LINECALC, TOUCHSTONE, SUPER COMPACT
 - + Non-Linear: LIBRA, HARMONICA
- ii) Time domain simulation:
 - + MICROWAVE SPICE

b) Graphic Tools:

- i) Layout: IC editor (ICED), MICAD on IBM PC's
- ii) Plotting: VALE on SUN workstations in conjunction with HP plotters

c) Photolithography Tool: David Mann 3000 mask generator 1/4 mil resolution.

All these tools were configured in a manner in which their outputs were as compatible as possible. Currently, circuit dimensions obtained from the simulation are still manually laid-out on ICED. Multilayer output CIF data are generated from ICED and subsequently are translated into MANN format to accommodate mask generation. For documentation purposes, these CIF data files can also be transported to the SUN networks and plotted on an HP plotter through a graphic package called VALE.

2) A fabrication procedure for thin-film MIM capacitors was investigated and verified. It features:

- a) Substrate: 10mil or 25mil alumina substrates with 1 µinch surface finish.
- b) First level metal: 1µm evaporated gold (Au) patterned using the lift-off technique. This includes the definition of the bottom

electrode of MIM capacitor and definition of inter-connecting microstrip lines.

c) Adhesion layer: 300 Å of TiW between adjacent metallic and dielectric layers to ensure structural integrity.

d) Dielectric layer: 2000 Å - 5000 Å of SiO₂ (Carleton University) and Si₃N₄ (CRC).

e) Top electrode and airbridge:

i) Windows are open where the top electrodes and pedestals of airbridge are to be situated.

ii) Definitions of top electrodes and airbridges are electroplated to 3µm-4µm and pattern-defined using the lift-off technique.

A surface finish of 1µinch or better was required for the MHMIC process because of the stringent requirements for MIM capacitor fabrication. Various studies indicated that 1µm first level metal was sufficient for microwave designs up to 26.5GHz. Only a slight improvement was observed when a thicker layer was used. Moreover, at 1µm, this layer of metal could be conveniently evaporated onto the substrate and lifted-off to define the pattern. The lift-off technique always yielded a more precise dimension than conventional etch-off techniques. This was because lifting off produced an A-shaped rather than a V-shaped cross-section as that obtained from latter techniques. Under a microscope, an A-shaped cross-section gives a true display of line width while a V-shaped cross-section does not [48].

An airbridge was proven to be a reliable DC and RF connection for the MHMIC process. Both MIM capacitors and spiral inductors were successfully fabricated using this technique. Thick electro-plated gold was used in this layer to leviate metallic losses as much as possible in the airbridge.

3) Measurement and characterization techniques for thin-film elements were devised and tested. The results are:

a) DC measurements (at 1MHz):

i) First batch: SiO₂ MIM capacitors

MIM SiO₂ capacitors, ranging from 0.2 pF - 50 pF were fabricated on 3 µinch surface finish substrates and DC tested. Errors on capacitance values were found to be as

large as 245%. Primarily, these alarming figures were due to the poor surface roughness of the substrates.

ii) Second batch: Si₃N₄ MIM capacitors

MIM Si₃N₄ capacitors, ranging from 0.7 pF - 20 pF, were fabricated on 1μinch surface finish to overcome the problem stated above. DC measurements indicated that the measured capacitance values agreed within 6.2% of the designed values.

b) S-parameter measurements:

The third batch of MIM Si₃N₄ capacitors, ranging from 0.1 pF - 20 pF were measured in the 0.5 GHz - 26.5 GHz frequency range. The TRL (Thru-Reflect-Line) technique was used to calibrate the HP8510 ANA and wafer prober set-up. Only the responses of the 0.5 pF, 3.9 pF and 20 pF capacitors were shown here. The RF responses of these 3 capacitors all indicated similar responses which exhibited the first self series resonance within the 0.5 GHz - 26.5 GHz frequency range. The resonant frequencies were inversely proportional to the capacitance values. They were approximately 24 GHz, 7 GHz and .3 GHz for the 0.5 pF, 3.9 pF and 20 pF MIM capacitor, respectively. Beyond these frequencies, the MIM capacitors behaved inductively. In any microwave design, the frequency of operation must therefore be less than the lowest self resonant frequency of all elements involved.

c) Physical modelling of MIM capacitors:

The proposed physical modelling technique was applied to the case of Si₃N₄ MIM capacitors, ranging from 0.1 pF - 20 pF. This modelling technique partitioned the MIM capacitor into many elementary microstrip discontinuities. These discontinuities were then modelled using lumped elements and electrically combined to simplify the equivalent circuit. The contributions of the respective microstrip discontinuities are listed in Tables 5.3, 5.4 and 5.5. The final MIM capacitor models have the equivalent circuits of that given in Figure 3.13 and Table 5.6.

This modelling technique differs from other modelling techniques at one crucial point: it requires no post-calculation optimization. The responses of the proposed models agree well

with the S-parameter measurements. Figures 5.6, 5.7 and 5.8 show the responses for the 0.5 pF, 3.9 pF and 20 pF, respectively. Also, on the same graphs, the Libra's TFC models were compared against the proposed models and the measurements. The comparison seemed to suggest that the Libra's TFC model can be further improved.

The physical modelling technique can also be applied to other microwave passive thin-film elements.

d) Resonant frequency shift (RFS) Q-factor measurements:

The RFS technique was applied to another batch of Si_3N_4 MIM capacitors, ranging from 0.6 pF - 23 pF. There were seven nearly square capacitors of values 0.6 pF, 1.5 pF, 3 pF, 5 pF, 11 pF, 17 pF and 23 pF in this batch. The fundamental unloaded resonant frequency of the single-section end-coupled linear resonator employed in this study was 2.85 GHz. In the 1 GHz - 40 GHz frequency range, available on the Wiltron 360 ANA, there were 14 harmonic responses of this resonator which could be observed. The DUT, which was a MIM capacitor, was loaded in series at the center of the resonator. The seven even harmonic responses indicated the unloaded characteristics of the resonator because they were not perturbed by the centrally loaded DUT. The seven odd harmonic responses, on the other hand, were directly perturbed by the DUT thus represent the loaded characteristics of the resonator. From the differences in resonant frequencies and Q-values between the unloaded- and loaded responses of the resonator, the characteristics such as resonance and Q-factors of the DUT were determined.

Figures 5.13, 5.14, 5.15 and 5.16 display the Q-responses of seven MIM capacitors under test as functions of frequency, in the 1 GHz - 40 GHz range. The results obtained from the RFS technique indicate the Q-factors are inversely proportional to capacitance and vary with frequency f according to $f^{(a-bf)}$, where a and b scalar coefficients. The measured Q-factor of the 0.2 pF MIM capacitor varied from 250 at 2 GHz to about 4 at 37 GHz while it varied from 60 at 2 GHz to 0.8 at 37 GHz for the 23 pF

case. The measured Q-factors of the other five MIM capacitors fell between these two boundaries.

The Q-responses of these seven MIM capacitors were also derived from their respective proposed physical models, under the assumption that the resistance parameter (see Figure 3.13) varied as a function of square root of frequency due to the skin-depth effect. These calculated Q-responses did not agree well with the measured results. Many authors have suggested that the resistive loss of a MIM capacitor should depend on frequency f according to $f^{(a-bf)}$. This justifies the fact that the RFS measurement technique is required to determine the Q-value of low-loss thin-film elements in general, and of MIM capacitors specifically.

6.2 Recommendations for Further Research:

In order to fully develop the MHMIC technology, other thin-film components such as spiral inductors and interdigital capacitors must also be characterized. The physical modelling and the RFS Q-measurement techniques are certainly applicable to such tasks.

The key issue in the physical modelling technique is the partitioning of the lumped components into adjoining microstrip elementary discontinuities. Once this critical step is successfully accomplished, the rest of the modelling exercise is very straightforward. It is to be noted that the lumped equivalents of these elementary microstrip discontinuities must finally be combined and simplified to a generally accepted and self-explanatory equivalent circuit.

In determining the Q-factors of thin-film components using the RFS technique, the frequency shifts can sometimes be so large that odd harmonics completely merge into their neighbouring even harmonics. This is the case for centrally loaded series connected spiral inductors and shunt connected capacitors. To overcome this obstacle, different connections or combined connections such as shunt connected spiral inductor, or LC in shunt must be attempted to lower the loading effects. Generally speaking, the type of connection can be conveniently selected to accommodate a particular perturbation requirement.

The TRL (Thru-Reflect-Line) calibration behaves satisfactory for the RFS measurement technique. However, its delay (line) standard must be selected in accordance with the frequency of interest. Many delay lines are thus required for a wideband coverage. The TAN (Thru-Alternate-Network) calibration will eliminate such inconvenience due to the inherent wideband characteristics of its standards.

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A I.4.1 Running ICED

After installing MICROWAVE_IC.BAT and .COM files, ICED is entered by:

```
CD ICED ! enter ICED directory
MICROWAVE_IC filename ! open filename.CEL
```

A I.4.2 Unnesting a file

When a drawing is completed, it may have a multi-level structure. This is possible because a file can call in other files, by adding new cells to its structure. This multi-level file must be flattened because the MANN generator does not have the capability to process a nested data file. This feature is available within ICED:

```
! Inside ICED already
_ SELECT CELL * ALL
_ UNGROUP
```

This procedure is to be repeated as many times as necessary to flatten all multi-level operations, rooted in the file. It is recommended that at least three trials are attempted. A useful check, for this is to exit ICED, then reopen the same file and observe to see whether any cell is called upon. If not, the file under consideration is effectively unnested.

A I.4.3 Creating a CIF File:

Once a layout is flattened, its CIF file can be generated for mask making and plotting purposes.

```
! Inside ICED already
_ CIF
_ SCALING FACTOR = 1
_ DONE
_ EXIT
A file named filename.CIF is created.
```

A I.4.4 Generating a MANN file:

Normalization of scaling factors and generation of MANN formatted text files are achieved by executing the steps as follows:

```
! Outside ICED, in DOS
_CIFLAT_W ! running CIFLAT_W.EXE
```

APPENDIX I: SIMULATION, LAYOUT AND DATA TRANSFERRAL INFORMATION

A I.1 Introduction

This appendix provides the information on microwave simulation, layout and computer data transferral required for the MHMIC process at Carleton University in 1988.

A I.2 Simulation and Optimization

After a preliminary manual design, computer simulation and optimization of the circuit are performed to converge to the design specifications. Simulation is also used to investigate circuit performance under various bias conditions or circuit sensitivity to component tolerance values, in either the frequency or the time domain. Problems such as out-of-band gain peak, oscillation, load pulling ... are often overlooked in the manual calculations. Moreover, accurate wideband responses are difficult to predict from manual calculations for multi-staged circuits. Computer simulation and optimization are a necessary step in the modern design process for microwave circuits.

A I.2.1 Software Capabilities:

Presently, TOUCHSTONE, LINECALC (trademarks of EEsof Ltd.) and SUPER COMPACT (trademark of Compact Software Ltd.) are available at Carleton University as CAD tools for microwave linear circuit analysis, circuit simulation and optimization, in the frequency domain. In the near future, LIBRA and HARMONICA will be obtained to perform the same tasks but for non-linear applications. SUPER COMPACT contains a module which analyzes and synthesizes basic microstrip, stripline and waveguide structures while TOUCHSTONE does not. LINECALC is introduced by EEsof to perform these functions. Thus, in general, TOUCHSTONE and LINECALC are to be purchased together to complete the set of linear circuit simulation software. With their versatile graphic capabilities and on-screen menu selection, these programs can greatly improve the efficiency of a designer. MICROWAVE SPICE (trademark of EEsof Ltd.) can also be used to analyze the circuit in the time domain if required. Detailed descriptions of these programs can be found in their operating manuals.

The operation of these programs are self-explanatory due to the on-

screen menu selection feature. However, efficient use of computer time, during optimization for example, requires also a high level of understanding of microwave design theories. Proper combination of computer power and theoretical knowledge prevents the waste of CPU time and allows the designer a high level of flexibility.

For either linear or nonlinear analysis, material and substrate parameters must be provided to the programs. These include:

a) Substrate and dielectric parameters

- relative dielectric constant
- thickness
- loss tangent
- purity
- surface roughness
- dispersion property

b) Metallization parameters

- thickness
- surface roughness
- losses (inductive and resistive)

c) Electrical models

- passive devices
- active devices
- interconnections and discontinuities

Output obtained, after running the simulations and optimizations, are dimensions, nodal connections and electrical parameters of the microwave circuit. The layout is then prepared according to these values.

A 1.2.2 Linear Simulation:

A 1.2.2.1 TOUCHSTONE and SUPER COMPACT

These are trademark products of EEsof Ltd. and Compact Software Ltd., respectively. They have similar capabilities and accuracy. They are supported on a variety of platforms:

- PC (IBM compatible)
- Apollo
- HP UNIX
- SUN UNIX
- VAX/VMS

Their functions are those of a standard analysis and simulation package. Basic features are:

i) Editor module: It is a simple, yet adequate, word processor which creates new and edits existing files.

ii) Analysis/Synthesis module: It is a numerical subroutine which calculates microwave responses of a linear circuit in the frequency domain.

iii) Tuner module: It is a subprogram which allows visual observation of performance improvement or degradation of a microwave circuit by altering the values of its parameters, one at a time.

iv) Optimizer/Yield module: It is a software module which interactively matches calculated output responses to those of a predetermined set of specifications, by optimizing the variable parameter's values. The object of this module is to minimize an error function which is defined by the numerical differences between the simulated responses and the desired goals.

v) Output module: It is a routine which produces plots and tables, in either magnitude/phase or Smith Chart format, of:

- S-parameters
- Noise information
- Stability information
- Impedance/admittance parameters

A I.2.2.2 LINECALC:

LINECALC is also a trademark product of EEsof Ltd. and is a module for studying the relationship between the physical attributes of a planar transmission line (eg. length, width, thickness ..) and its electrical parameters (eg. electrical length, characteristic impedance, loss ...).

LINECALC is intended to complement TOUCHSTONE, on both PC and mainframe environments.

A I.2.3 Nonlinear simulation:

A I.2.3.1 LIBRA and HARMONICA:

These competitive, yet similar in capabilities, programs offer both time and frequency domain analysis of nonlinear circuits (oscillators, multipliers, mixers, power amplifiers...). Presently, only HARMONICA offers nonlinear circuit optimization. However, in the near future, automatic optimization against DC bias settings, input power levels... will be available on both

programs.

LIBRA and HARMONICA employ the modified harmonic balance method [12], [13] to analyze the steady-state circuit response. In this method, the circuit is divided into a linear subcircuit and a nonlinear subcircuit. The linear subcircuit is analyzed in the frequency domain whereas the nonlinear section is analyzed in the time domain. At the interface between the subcircuits, the complex coefficients of the current and voltage signals are balanced at the point of convergence. The transformations between frequency and time domains are performed via the Fast Fourier Transform (FFT) algorithm. For reasonable convergence times, the number of signal harmonics is usually limited to less than ten.

A I.2.3.2 Microwave SPICE:

This program performs time domain analysis of nonlinear circuits. It is more accurate than LIBRA or HARMONICA but not as fast especially for circuits with long time constant. However, it gives the transient response, which may be more important in studies such as the start-up of oscillations in microwave oscillators.

A I.3 Circuit layout

At Carleton University, IC editor (ICED) is employed as a layout package for mask making applications since it can generate Caltech intermediate format (CIF) code and a program for CIF to MANN format conversion is available. Other packages such as MICAD or VALE can also be used but another conversion program is required to obtain the MANN format.

ICED and MICAD run on IBM PCs while VALE is dedicated to the UNIX environment such as provided on SUN workstations. The sequence, used to invoke each program, will be described in detail in Section A I.6.

It is required that the following information be available prior to drawing the circuit:

a) Fabrication rules and guidelines:

- undercutting and misalignment compensations
- mask generator and process resolutions
- format translation restrictions.

b) Electrical limitations:

- allowable dimensions and parameter values for elements
- circuit's configurations (with or without via holes, wafer probe or SMA connect...)

Hard copies of the drawing are provided with clarity and quality, for illustrative purposes, by ICED via the HP7580 plotter. Also, graphic data output, although not fully, is adequately compatible to that of the mask generator, which utilizes the MANN format.

A I.3.1 ICED

ICED is a multi-purpose, multi-layer graphic package, which has been custom tailored, by the author, to facilitate the layout of a microwave circuit. Colors and names of the layers are chosen to portray a realistic unambiguous picture of the designed circuit. They are as follows:

Table A I.1

Layer Definitions of Thin-film Microwave Process on ICED

DESCRIPTION	NAME	COLOUR
Resistive pattern	RES	Amber
Conductor	CON	Solid yellow
Bottom electrode window	CAP	Outlined yellow
Dielectric pattern	DIE	Sky blue
Top electrode or airbridge pedestal	PED	Navy blue
Airbridge	AIR	Green

The smallest unit on ICED is chosen to be 0.25 mili-inch (or mil) which is the resolution of the DAVID MANN 1600 mask generator. Hence, all layouts, on screen, are of ratio 1:1 to the actual size of the circuit. The scalings of ICED's numerical output data, will be described in detail in Section A I.4.

ICED allows definitions of polygons, boxes and wires to be used in a circuit layout. One major drawback that is ICED only permits angles which are multiples of 45 degrees. Thus, irregular shapes are not supported on an ICED pattern. For example, a MIM capacitor is laid out as follows:

Figure A I.1 Layout of a MIM capacitor using ICED

The area of the top electrode determines the capacitance of the overlay structure, which contains a fixed-thickness dielectric layer. The dielectric layer is intentionally drawn larger than both electrodes to eliminate shorting the two plates together, in case of fabrication misalignment. The pedestals, or sometimes known as the contact windows, are open on the top electrode and output line to begin the construction of the airbridge. The airbridge is electroplated up from the contact windows to form a permanent connection.

As shown and described, every physical and electrical connection of an element, or a circuit, is truly realized and clearly illustrated by ICED, in a multi-layer environment. Additionally, ICED is relatively inexpensive and very user-friendly. These features make ICED an attractive layout package.

However, when using ICED, one major drawback is found. The present in-house conversion program which translates the CIF files, generated by ICED, to the MANN format, acceptable to the mask generator, allows only operations involving rectangular boxes. Operations such as rotations, inclinations ... will cause dimension round-offs and/or coordinate shiftings in the output data file, due to the format-translation process. These problems are common to all format conversion programs. The errors may cause the masks generated to have disjoint or disfigured patterns. This is why the MIM capacitor, shown above, is comprised only of rectangular boxes. Also the dimensions of a rectangular box must be integer, greater than one, multiples of the pattern generator's aperture resolution, in order for the coordinates of its center to be integers. Otherwise, these coordinates will be shifted by the translation programs or rejected by the mask generator.

Format translation difficulties and the pattern generator's resolution place severe limitations on the layout process. Layout guidelines have to be strictly obeyed to avoid unnecessary errors. At present, the performance of these translation programs are adequate, however in the future, they have to be upgraded to fully utilize ICED. MASKAP (trademark of Pheonix Software Inc.) is one possible replacement translator, which is commercially available. Otherwise, a new layout package, which directly outputs MANN formatted data, has to be found to solve the problems of incompatibility.

A I.3.2 MICAD

MICAD is a multilayer graphic package developed by EEsof. It allows operations involving arbitrary angles. Hence, most shapes and patterns are accepted. Moreover, MICAD and other utilities from EEsof Ltd. can also be linked to TOUCHSTONE by ACADEMY, a microwave circuit design environment, to further automate the design process of a microwave circuit. After running TOUCHSTONE, the simulated information can be transferred to MICAD for either electrical or physical realization of the circuit. They are laid out automatically via ACADEMY. Indeed, undesirable arrangements of circuit elements, in the automatic drawing, exist and must be manually corrected.

Unfortunately, this graphic package is not suitable for the thin-film process at Carleton because within MICAD only one layer can be visually inspected at a time and MICAD does not provide CIF output. It is capable of providing GDS II, CALMA and HPGL outputs which are not directly compatible with the MANN mask generator.

A I.3.3 VALE:

VALE is available on the METHEUS, a peripheral of the SUN network. This multilayer package is reconfigured, by Dr. M. Lefebvre, to support layout of a microwave thin-film circuit. However, since the METHEUS is dedicated to designers of the silicon VLSI group, it is only used here for plotting purposes. One version of VALE has been modified to facilitate six layers of microwave thin-film layout. CIF films generated by ICED can be transformed into PHL format to be read by VALE. Hard copies of these plots can be then obtained via the HP7500 plotter. Detailed description of the CIF-to-PHL format translation is explained in Section A I.5.

A I.4 Maskmaking: CIF to MANN Format Conversion

CIF files, generated by ICED, must be converted to MANN format, in order to be accepted by the DAVID MANN 1600 mask generator. Three Fortran programs, written by D. Beard at Carleton University, are available for this purpose. They are:

- i) CIFLAT_W: to unnest the CIF data file
- ii) CONVERT: to reorganize the CIF file
- iii) CIF2MAN: to translate from CIF format

In the MICROWAVE_IC.BAT and .COM files, layer definitions and screen resolution are set. Layer definitions are those provided in sections A I.3.1. The resolution is set at 0.25 mili-inch (mil), which is that of the MANN mask generator. Hence, all layouts on screen are of ratio 1:1 with respect to the dimensions of the circuit. The scaling factor must be set to one, when a CIF file is created from ICED, and the internal elementary unit of ICED, lambda, must be set to 500 micro-inch, when running CIF2MAN, so that all dimensions are properly conserved. Scaling factors, embedded within CIFLAT_W, CONVERT and CIF2MAN, are chosen accordingly to ensure conservation of dimensions. Summary of the scaling process, which occurs during format translation, is as follows:

Table A I.2

Scaling Factors in Data Conversion from CIF to MANN Format

Location of occurrence	Internal scaling factor (set by program)	External scaling factor (set by operator)
Generating CIF file from ICED	4	1
Running CIFLAT_W and CONVERT	1/4	NA
Running CIF2MAN	1/2*	2*

* NOTE: When running CIF2MAN, lambda is set to 500 micro-inch (0.5 mil), which is the minimum allowable entry. This is twice as large as the desired resolution of 0.25 mil. Thus, CIF2MAN must internally provide a scaling factor of 1/2. As a result, the actual lambda is 0.25 mil.

A I.4.1 Running ICED

After installing MICROWAVE_IC.BAT and .COM files, ICED is entered by:

```
CD ICED ! enter ICED directory
MICROWAVE_IC filename ! open filename.CEL
```

A I.4.2 Unnesting a file

When a drawing is completed, it may have a multi-level structure. This is possible because a file can call in other files, by adding new cells to its structure. This multi-level file must be flattened because the MANN generator does not have the capability to process a nested data file. This feature is available within ICED:

```
! Inside ICED already
_ SELECT CELL * ALL
_ UNGROUP
```

This procedure is to be repeated as many times as necessary to flatten all multi-level operations, rooted in the file. It is recommended that at least three trials are attempted. A useful check, for this is to exit ICED, then reopen the same file and observe to see whether any cell is called upon. If not, the file under consideration is effectively unnested.

A I.4.3 Creating a CIF File:

Once a layout is flattened, its CIF file can be generated for mask making and plotting purposes.

```
! Inside ICED already
_ CIF
_ SCALING FACTOR = 1
_ DONE
_ EXIT
```

A file named filename.CIF is created.

A I.4.4 Generating a MANN file:

Normalization of scaling factors and generation of MANN formatted text files are achieved by executing the steps as follows:

```
! Outside ICED, in DOS
_CIFLAT_W ! running CIFLAT_W.EXE
```

```

+INPUT FILE NAME =  A:filename.CIF
+OUTPUT FILE NAME = A:filename.FLT
_CONVERT          ! running CONVERT.EXE
+INPUT FILE NAME =  A:filename.FLT
+OUTPUT FILE NAME = A:FILENAME.CON
_CIF2MAN          ! running CIF2MAN.EXE
+INPUT FILE NAME =  A:filename.CON
+OUTPUT PREFIX =    A:XXXXXX ! 5 letters only
+COMMENTS =         Date, filename, lambda value...
+LAMBDA =            500 ! in micro-inch
+FIDUCIAL MARK =    NO  ! if no step-and-repeat
+REFLECT ON X-AXIS:
    *YES            ! for dark field mask
    *NO             ! for light field mask

```

Six text files are generated. Each contains architectural information of a different layer. They are:

```

_A:XXXXXX_ES.TEX    ! resistive layer
_A:XXXXXX_ON.TEX    ! conductive layer
_A:XXXXXX_AP.TEX    ! bottom electrode layer
_A:XXXXXX_IE.TEX    ! dielectric layer
_A:XXXXXX_ED.TEX    ! pedestal (or top electrode layer)
_A:XXXXXX_IR.TEX    ! air bridge layer

```

The syntax of these text files can be checked by running VRFYMANN A:XXXXXX_XX.TEX. After this, they are ready for the DAVID MANN photomask generator.

It must be noted that the new version of ICED introduces a line which displays "; 9 filename" at the beginning of the CIF file. This line must be removed in order for CIFLAT_W to run properly.

A 1.5 Plotting: CIF to PHL format conversion

A microwave technology file is created within VALE, on the METHEUS, to process CIF files transferred from the PC to the SUN network. This technology file must be installed and invoked in the system set-up file of a SUN account, by Dr. M. Lefebvre. In order to be accepted by VALE, these CIF files must be corrected in format and then converted to PHL. VALE accepts PHL formatted files to create plotting windows on the screen. A specific

device or connection can be zoomed in, on a particular window, to reveal all features. A hard copy, of each window, is obtained via the HP7580 plotter.

A I.5.1 Converting from PC-CIF to SUN-CIF format

The CIF format, produced from ICED, is not the same as its counterpart on the SUN. This type of discrepancy is rather common in the world of computer design. Thus, the CIF files, generated on the IC, must be corrected in format, before they can be accepted by the SUN. The correction of the ASCII file can be performed using any word processor such as WORDPERFECT or WORDSTAR. Necessary spaces, or blanks, must be inserted as shown below:

! After creating a CIF file in ICED

PC-CIF Format	SUN-CIF Format
LXXX	L XXX
TXXX	T XXX
CX	C X

where: X = any character or number

A I.5.2 Converting from CIF to PHL format on the SUN

PHL is one of many graphic formats, which the SUN supports, required when using VALE. Program CIF2PHL is available on the SUN to translate SUN-CIF files to PHL format.

! On the METHEUS:

_CIF2PHL filename.CIF ! converting filename.CIF to
filename.PHL

_VALE filename ! open filename.PHL in VALE

_YANK ! extract image from drawing
on screen

_YSAVE new_filename ! create new window to store image
! just extracted

_:q ! quit VALE

_PHLPLOT new_filename ! plot file using the HP7580

A I.6 Data transferral

Previous sections described the translation processes among one data

format to others, to facilitate specific tasks such as mask making and plotting. Within the process, transferring files between computer systems is unavoidable. Fortunately, the SUN and the PCs are networked together making file transferral simple and efficient. Transfer of files between the PC and the SUN can be done on the PC via KERMIT.

A 1.6.1 From PC to SUN:

The following sequence is to be invoked when sending a file from a PC, which has been hardwired to the SUN network.

```
! On a PC connected to the SUN
! In a drive which contains KERMIT.EXE
_KERMIT          ! on PC KERMIT
_CONNECT         ! connect to SUN network, using PC screen as a
                  SUN's terminal
    LOGIN        ! Login to SUN account
    PASSWORD
_KERMIT          ! run SUN KERMIT
_RECEIVE         ! set SUN account to RECEIVE mode
_<CTRL>]C        ! return to PC KERMIT to send file
_SET BAUD 9600   ! set data rate
_SET PORT 2      ! and output port
_SEND filename.CIF ! send CIF file to SUN account
! when sending is DONE
_CONNECT        ! at back to SUN account
_LS            ! list current account to make sure file
               transferring is successful
_QUIT          ! if OK, exit SUN KERMIT; if not, try again
_<CTRL>D        ! logout of SUN account
_<CTRL>]C       ! return to PC
_QUIT          ! exit PC KERMIT
```

A 1.6.2 From SUN to PC:

The following sequence should be invoked when sending a file from the SUN network to a PC.

```
! On a PC connected to the SUN
! In a drive which contains KERMIT.EXE
```

```

_KERMIT          ! run PC KERMIT
_SET BAUD 9600    ! define data rate
_SET PORT 2       ! and output port
_RECEIVE         ! set PC drive in RECEIVE mode
_CONNECT         ! connect to SUN network using PC screen as a
                  SUN's terminal
    LOGIN        ! Login to SUN account
    PASSWORD
_KERMIT          ! run SUN KERMIT
_SEND filename   ! send filename to PC
! when receiving is DONE
_<CTRL>]C        ! return to PC
_DIR A:          ! check
_CONNECT         ! if OK, get back to SUN account, if not, try again
_QUIT           ! exit SUN KERMIT
_<CTRL>D        ! logout of SUN account
_QUIT           ! exit PC KERMIT

```

A I.6.3 Between HP and PC:

With the use of the HP82321 interface card and Hewlett Packard (HP) translation software, HP equipment can be networked and controlled with an IMB PC.

The HP interface bus (HPIB) is available at the I/O port of the interface card. Data obtained from HP instruments are conveniently translated to DOS formatted data files, through the HPIB, via the utility HP_UTILS, to be accessed by TOUCHSTONE or SUPER COMPACT later on. Basic functions of this process are discussed here. For in-depth understanding, one should consult the manual of the VECTRA PC, marketed by HP.

There are three formats supported in this scheme: DOS, LIF and HPW.

HP-LIF files are HP-BASIC formatted files. LIF files can only be read, edited, run and stored inside HP-BASIC. LIF formatted discs are required to store LIF files. Since the hard disc of the IC is already configured in DOS format, only 3 1/2 inch floppy discs are acceptable media to store LIF files.

HP-HPW files are of an intermediate format between DOS and LIF. Sequentially, HP-DOS files can be translated to HP-HPW, then to HP-LIF files, and vice versa. HPW files are stored inside the HPW-DIR directory, in the

DOS environment. This directory is automatically created by HP-UTILS, during installation, to separate DOS from HPW files. It is the HPW files which makes the translation between LIF and DOS formats possible.

A I.7 Summary

The design process for a microwave thin film circuit has been described. Appropriate numerical simulation and graphic layout software packages are selected to facilitate the design process. The simulations of linear or nonlinear circuits are performed using existing microwave CAD packages, commercialized by EEsof and Compact Software. The layout of the physical circuit is drawn on ICED, using information provided by the simulations. CIF files, generated from ICED, are converted to MANN format for mask making and to PHL format for plotting purposes. Layout guidelines and restrictions are clearly listed to ensure complete compatibility among translatable data formats so that all dimensions are properly conserved.

Due to difficulties in translating CIF to MANN format, circuits can be drawn only using rectangular boxes. Also, only angles of 0, 90, 180, 270 degrees are allowed. These place severe limitations on the layout process. In the future, better translation software, such as MASKCAP, or MANN-format-output graphic package should be employed to eliminate these problems.

Data transfer between HP instruments and the PC is accomplished via the HP82321 interface card and HP translation software. Conveniently, HP measured data files can be converted to DOS format for simulations in TOUCHSTONE or SUPERCOMPACT.

APPENDIX II

CRC Thin-film MHMIC Process Description (TiW resistors, Si₃N₄ MIM capacitors, Air bridges)

This appendix only describes the processing steps which are different from those used by the Carleton University thin-film fabrication process. Unless stated otherwise, steps such as CLEAN and LIFT-OFF are similar to both processes and can be referenced back to the report written by B.A. Syrett and L.P. Berndt [8].

- 1) **First level metal (lift-off or reverse process):** light field mask
 - 1.1) Spin coat photoresist (5214) onto substrate at 3000 RPM
(bake at 90 degrees for 15 minutes)
 - 1.2) Expose under mask aligner (FLM mask) for 2 seconds
 - 1.3) Bake on hot plate at 140 degrees for 40 seconds
 - 1.4) Develop pattern using 400K solution (4 part water : 1 part solution)
for 40 seconds
 - 1.5) Plasma etch (de-scum) in O₂ for 30 seconds
 - 1.6) Degrease in NH₄OH for 20 seconds
 - 1.7) Evaporate Ti (300Å), Pt (1000Å), Au (10000Å)
 - 1.8) Lift-off to define pattern
- 2) **Resistor level:** light field mask
 - 2.1) Spin coat photoresist (1370) onto substrate at 5000 RPM
(bake at 90 degrees for 15 minutes)
 - 2.2) Expose under mask aligner (RES mask) for 5 seconds
 - 2.3) Develop pattern using the same developer as in step 1.4 for 45
seconds
 - 2.4) Bake in oven at 120 degrees for 15 minutes
 - 2.5) Etch TiW using H₂O₂
 - 2.6) Strip photoresist at 90 degrees
 - 2.7) Clean substrate
 - 2.8) Bake in the 400-425 degree range to achieve required value of sheet
resistance
- 3) **Dielectric level:** light field mask
 - 3.1) Plasma etch (de-scum) in O₂ for 30 seconds
 - 3.2) Degrease in NH₄OH for 20 seconds
 - 3.3) Deposit 2000-4000Å of Si₃N₄
 - 3.4) Spin coat photoresist (1370) onto substrate at 5000 RPM
(bake at 90 degrees for 15 minutes)
 - 3.5) Expose under mask aligner (DIE mask) for 5 second
 - 3.6) Develop photoresist for 45 seconds
 - 3.7) Bake in oven at 120 degrees for 15 minutes

- 3.8) Plasma etch in CF_4 (approximately 5 minutes for a $1 \times 1 \text{ in}^2$ substrate with $2000\text{\AA}$ dielectric thickness)
- 3.9) Strip photoresist at 90 degrees
- 3.10) Clean substrate
- 3.11) Plasma etch (de-scum) in O_2 for 30 seconds
- 3.12) Degrease in NH_4OH for 20 seconds

4) Air bridge pedestal, via hole pad or MIM top electrode level: dark field mask

- 4.1) Spin coat photoresist (1350) on to substrate at 5000 RPM (bake at 90 degrees for 15 minutes)
- 4.2) Expose under mask aligner (WIN mask) for 5 seconds
- 4.3) Develop photoresist for 45 seconds
- 4.4) Bake in oven at 120 degrees for 15 minutes
- 4.5) Plasma etch (de-scum) in O_2 for 30 seconds
- 4.6) Degrease in NH_4OH for 20 seconds
- 4.7) Evaporate Ti ($300\text{\AA}$) and Au ($4000\text{\AA}$)

5) Air bridge span level: dark field mask

- 5.1) Spin coat adhesion promoter (MS804) onto substrate at 5000 RPM (bake at 90 degrees for 5 minutes)
- 5.2) Spin coat photoresist (1350) onto substrate at 2000 RPM (bake at 90 degrees for 15 minutes)
- 5.3) Repeat step 5.2
- 5.4) Expose under mask aligner (ABR mask) for 15 seconds
- 5.5) Bake in oven at 90 degrees for 15 minutes
- 5.6) Develop photoresist using (3 part water : 1 part solution) 400K developer
- 5.7) Bake in oven at 90 degrees for 15 minutes
- 5.8) Electroplate in gold solution to get $4\text{--}5 \mu\text{m}$ Au thickness
- 5.9) Develop photoresist using (4 part water : 1 part solution) 400K developer
- 5.10) Develop photoresist using (3 part water : 1 part solution) 400K developer
- 5.11) Etch Au seed layer at 50 degrees (enough to remove $4000\text{\AA}$ Au)
- 5.12) Etch Ti seed layer at 25 degrees for 5 seconds
- 5.13) Plasma etch (de-scum) in O_2 for 30 seconds
- 5.14) Strip photoresist at 90 degrees using only 5% strength solution
- 5.15) Clean substrate

APPENDIX III

! Microwave technology file on ICED

!

! Define display and plot data format

View Off

Menu Menu

View Limit on Scale=0.5 Depth=1

Snap=1.0

Plot Plotter HP7580

!

! Define grid

! Five red ticks between adjacent magenta ticks,

! Two magenta ticks between adjacent cyan ticks

Grid 1 On Step=1 Red

Grid 2 On Step=5 Magenta

Grid 3 On Step=2 Cyan

!

! Define names and colours for layers

!

! Default layer

Layer * Solid Yellow

!

! Resistive layer

Layer 1 Name="RES" Solid Magenta CIF="RES" Pen=3

!

! Metal (gold) layer

Layer 2 Name="CON" Solid Yellow CIF="CON" Pen=8

!

! Seed layer for MIM capacitor

Layer 3 Name="CAP" Dotted Yellow CIF="CAP" Pen=7

!

! Dielectric layer

Layer 4 Name="DIEL" Solid Cyan CIF="DIE" Pen=5

!

! Airbridge pedestal (or MIM top electrode) layer

Layer 5 Name="PEDE" Solid Blue CIF="PED" Pen=2

!

! Airbridge layer

Layer 6 Name="AIRB" Solid Green CIF="AIR" Pen=4

!

! Text layer

Layer 100 Name="TEXT" Solid White CIF=" " Pen=1

APPENDIX IV

PART A: Sample calculations for the 3.9 pF MIM capacitor (see Figure 5.7)

(* Calculate the main, stray and coupling capacitances of MIM structures *)

(* *)

diethick=0.2 ; (* in micron *)

erdie=7, (* Er of silicon nitride *)

metalthick=1 ; (* in micron *)

capwidth=109; (* width of top electrode, in micron *)

caplength=109; (* in micron *)

(*) (*)

(* Calculate DC capacitance C1 *)

$$C1 = \epsilon_r \epsilon_0 * (8.854 * 10^{(-6)}) * \text{capwidth} * \text{caplength} / \text{diethick}; \quad (* \text{ in pF} *)$$

```
Print[N[C1]];
```

(* *)

gap=0.63; (* gap between bottom electrode
and output feedline, in mil *)

ersub=9.8 ; (* Er of Alumina substrate *)

subheight=10; (* in mil *)

outputwidth=2.91 ; (* width of output feedline, in mil *)

bottomwidth=4.65; (* width of bottom electrode, in mil *)

(*) (*)

(*) Coupling between input and output ports is C2,

Stray capacitance at input port is C_3 ,

Stray capacitance at output port is C4 *)

(* *)

(* Due to microstrip gap *)

```
smallwidth=Min[outputwidth,bottomwidth];
```

```
ratio1=smallwidth/subheight ;
```

$$\text{Modd} = \text{ratio1} * (0.619 * \text{Log}[10, \text{ratio1}] - 0.3853) ;$$

Kodd=4.26 - 1.453*Log[10, ratio1] ;

```
ratio2=gap/smallwidth;
```

If[$\text{ratio2} < 0.3$,

Meyen=0.8675 ;

$$\text{Keven} = 2.043 \cdot \text{ratio1}^{0.12} ;,$$
$$\text{Meven} = (1.565 / \text{ratio1}^{0.16}) - 1 ;$$

```
Keven=1.97 - 0.03/ratio1 ;]
```

$$\text{Codd} = \text{smallwidth} * 2.54 * 10^{(-5)} * (\text{ratio2}^{\text{Modd}}) * \text{Exp}[\text{Kodd}] * 9.6 * (\text{ersub}/9.6)^{0.8};$$
$$\text{Ceven} = \text{smallwidth} * 2.54 * 10^{(-5)} * (\text{ratio2}^{\text{Meven}}) * \text{Exp}[\text{Keven}]$$
$$*9.6*(ersub/9.6)^{0.9} ;$$

```

Print[N[Codd]];
Print[N[Ceven]];
(* Calculate coupling capacitance C2*)
C2=0.5*(Codd - 0.5*Ceven);    (* in pF *)
Print[N[C2]];
C3_gap=0.5*Ceven;              (* in pF *)
C4_gap=0.5*Ceven;              (* in pF *)
(* Due to microstrip steps *)
(* *)
wsmallin=74;    (* width of small line at input port, in micron *)
wbegin=118;     (* width of big line at input port, in micron *)
wsmallout=10;   (* width of small line at output port, in micron *)
wbigout=74;     (* width of big line at output port, in micron *)
ratio3=wbegin/wsmallin;
product3=wbegin*wsmallin*10^(-12);
If[ratio3<1.5,
  C3=C3_gap,
  If[ratio3>3.5,
    (* total C3 is sum of stray capacitances due to gap and step *)
    C3=C3_gap+(product3^0.5)*(130*Log[10,ratio3] - 44),
    C3=C3_gap+(product3^0.5)*((10.1*Log[10,ersub]+2.33)*ratio3
      - 12.6*Log[10,ersub] - 3.17)]];
Print[N[C3]];
ratio4=wbigout/wsmallout;
product4=wbigout*wsmallout*10^(-12);
If[ratio4<1.5,
  C4=C4_gap,
  If[ratio4>3.5,
    (* total C4 is sum of stray capacitances due to gap and step *)
    C4=C4_gap+(product4^0.5)*(130*Log[10,ratio4] - 44),
    C4=C4_gap+(product4^0.5)*((10.1*Log[10,ersub]+2.33)*ratio4
      - 12.6*Log[10,ersub] - 3.17)]];
Print[N[C4]];
(* *)

```

PART B: Sample calculations for the 3.9 pF MIM capacitor (see Figure 5.7)

(* Calculate the series inductance and resistance of MIM structures *)

(* *)

t=1; (* metal thickness *)

h=0.000635; (* substrate height *)

```

(* *)
(* Define input data matrix
    dim1[[1,1]]=length
    dim1[[1,2]]=width *)
dim1={{16,10},      (* length and width of airbridge *)
      {129,118}};  (* length and width of one electrode, use worst case *)
Array[a,2];         (* array of calculated inductances *)
Array[c,2];         (* array of calculated resistances *)
(* *)
(* Calculate inductance and resistance due to microstrip lines *)
For[i=1,i<3,i++,
  l=dim1[[i,1]];
  w=dim1[[i,2]];
  ratio1=l/(w+t);
  term1=Log[ratio1];
  term2=1.193;
  term3=0.2235*(1/ratio1);
  a[i]=0.0002*l*(term1+term2+term3)*(0.57-0.145*Log[w/(1000000*h)]);
  c[i]=0.029*ratio1/2;
];
(* Calculate inductances introduce by the steps, for Wbig/Wsmall<5.0
    dim2[[1,1]]=large width
    dim2[[1,2]]=small width *)
dim2={{118,74},    (* microstrip step at input port *)
      {74,10}};    (* microstrip step at output port *)
Array[b,2];
For[i=1,i<3,i++,
  wbig=dim2[[i,1]];
  wsmall=dim2[[i,2]];
  ratio2=wbig/wsmall;
  b[i]=h*(40.5*(ratio2-1)-32.57*Log[ratio2]+0.2*(ratio2-1)^2)/(Log[ratio2]);
];
(* *)
(* Calculate total inductance and total resistance *)
Print[N[ind=a[1]+2*a[2]+b[1]+b[2]]];
Print[N[res=c[1]+2*c[2]]];

```

APPENDIX V

(* Sample calculations for the 0.6 pF MIM capacitor *)

(* *)

(* Input data array: freqBW

frequency array (fo, f2, f1) obtained from Wiltron 360

+ centre frequency: fo

+ 3-dB bandwidth: f2-f1

odd rows are odd harmonic (loaded) measurements

even rows are even harmonic (unloaded) measurements *)

(* *)

```
freqBW=({4.084,4.060,4.108},  
        {5.640,5.615,5.669},  
        {9.060,9.021,9.095},  
        {11.224,11.186,11.262},  
        {14.34,14.292,14.388},  
        {16.720,16.672,16.772},  
        {19.656,19.593,19.718},  
        {22.136,22.072,22.199},  
        {24.932,24.852,25.011},  
        {27.472,27.394,27.548},  
        {30.184,30.108,30.317},  
        {32.788,32.709,32.964},  
        {35.356,35.212,35.487},  
        {37.876,37.670,37.968}});
```

(* *)

(* input data array: mag

S21(magnitude_dB) obtained from Wiltron 360 *)

(* *)

```
mag={(-33.723},  
     {-26.309},  
     {-22.148},  
     {-18.919},  
     {-17.008},  
     {-14.862},  
     {-15.231},  
     {-13.008},  
     {-12.812},  
     {-12.926},  
     {-12.932},
```



```

        {-12.143},
        {-10.406},
        {-10.800});
numfreq=14;
(* *)
(* Calculate Q-values, resonant frequencies and coupling factors *)
(* *)
(* Q-value array *)
Array[qmeas,numfreq];
(* Resonant frequency array *)
Array[fmeas,numfreq];
(* Beta value (coupling factor) array *)
Array[b,numfreq];
For[i=1,i<(numfreq+1),i++,
    f1=freqBW[[i,3]];
    f2=freqBW[[i,2]];
    (* *)
    fmeas[i]=freqBW[[i,1]];
    q0=fmeas[i]/((f1-f2));
    t=10^(mag[[i]]/20);
    b[i]=t/(2*(1-t));
    qmeas[i]=q0*(1+2*b[i]);
    Print[N[b[i]]];
    Print[N[qmeas[i]]];
    Print[N[fmeas[i]]];
];
(* *)
(* Calculate loaded and unloaded responses at odd-order frequencies*)
(* *)
(* Even-order frequency array *)
Array[freqeven,numfreq];
(* Odd-order frequency array *)
Array[freqodd,numfreq];
(* Even-order Q array *)
Array[qeven,numfreq];
(* Odd-order Q array *)
Array[qodd,numfreq];
(* Separate odd from even responses *)
For[i=1,i<(numfreq+1),i++,
    If[EvenQ[i],

```

```

        freqeven[i/2]=fmeas[i];
        qeven[i/2]=qmeas[i];
        ieven=i/2;;
        freqodd[(i+1)/2]=fmeas[i];
        qodd[(i+1)/2]=qmeas[i];
        iodd=(i+1)/2;
    ]
]
(* Fit low even-order (unloaded) responses to a linear log-log line also *)
QTable2=Table[Flatten[{Log[freqeven[j]],Log[qeven[j]]}],{j,1,(ieven-3),1}];
Qeven[k_]=Fit[QTable2,{1,k},k];
Print[Qeven[k]];
Show[ListPlot[QTable2],ParametricPlot[{k,Qeven[k]},{k,1,5}]];
(* Interpolate even-order responses to obtain unloaded responses at odd-order
frequencies *)
Array[freqsys,iodd];
Array[qsys,iodd];
Array[freqzero,iodd];
Array[qzero,iodd];
avg=(freqeven[ieven]-freqeven[1])/(ieven-1);
For[g=1,g<(iodd+1),g++,
    freqsys[g]=freqodd[g];
    qsys[g]=qodd[g];
    If[g==1,freqzero[1]=freqeven[1]/2,
    If[LogicalExpand[Not[g==1]&&Not[g==iodd]],
        freqzero[g]=(freqeven[g-1]+freqeven[g])/2,
    If[LogicalExpand[(g==iodd)&&(iodd>ieven)],
        freqzero[iodd]=freqeven[ieven]+ freqeven[ieven]/(2*ieven),
        freqzero[iodd]=(freqeven[ieven-1]+freqeven[ieven])/2
    ]];
    qzero[g]=Exp[Qeven[Log[freqzero[g]]]];
Print[freqsys[g]];
Print[freqzero[g]];
Print[qsys[g]];
Print[qzero[g]];
]
(* *)
(* Determine Q-values from differences between loaded and unloaded
responses at odd-order frequencies *)
(* *)
Array[freq,{iodd,2}];

```

```

Array[qvalue,iodd,2]);
For[i=1,i<(iodd+1),i++,
    freq[i,1]=freqsys[i];
    freq[i,2]=freqzero[i];
    qvalue[i,1]=qsys[i];
    qvalue[i,2]=qzero[i];
]
Array[theta1,iodd];
Array[reactance,iodd];
Array[Qmim,iodd];
Array[Rmim,iodd];
(* Calculate DUT reactances at odd-order frequencies *)
For[i=1,i<(iodd+1),i++,
    n=2*i-1;
    fsys=freq[i,1]*10^9;
    fzero=freq[1,2]*10^9;
    fratio=fsys/fzero;
    deltaf=freq[i,1]-freq[i,2];
    (* *)
    theta1[i]=(*fsys*Pi/fzero*) (n+deltaf/freq[1,2])*Pi;
    reactance[i]=100*Cot[theta1[i]/2];
];
(* Fit DUT reactance to a function of the form (af+b/f), where f is frequency *)
XTable=Table[{freq[i,1],N[reactance[i]]},{i,1,iodd,1}];
Xfunc[f]=Fit[XTable,{1/f,i},f];
Print[XTable];
Print[Xfunc[f]];
(* Calculate DUT series inductance and series capacitance from fitted reactance curve *)
If[LogicalExpand[(Coefficient[f*Xfunc[f],f,2]<0)],
    ind=0,
    ind=Coefficient[f*Xfunc[f],f,2]/(2*Pi*10^9)];
cap=(-1)/(Coefficient[f*Xfunc[f],f,0]*2*Pi*10^9);
Print[N[ind]];
Print[N[cap]];
(* Calculate energies stored in linear resonator alone, in DUT and in linear resonator loaded by DUT *)
For[i=1,i<(iodd+1),i++,
    (* Energies stored in linear resonator *)
    E0=50*(2*theta1[i]-Sin[2*theta1[i]])/(16*Pi*freq[i,1]*10^9);

```

```

Print[N[E0]];
(* Energies stored in DUT, which is the 0.6 pF MIM capacitor *)
Ec=(Sin[theta1[i]/2])^2/(4*(2*Pi*freq[i,1]*10^9)^2*cap)+
      (Sin[theta1[i]/2])^2*ind/4;
Print[N[Ec]];
(* Energies stored in entire system *)
Print[N[Ec+E0]];
(* Calculate Q-values of DUT *)
Qmim[i]=N[Ec/((E0+Ec)/qvalue[i,1]-E0/qvalue[i,2])];
Print[N[Qmim[i]]];
];
(* Plot Q-factor versus frequency *)
Qgraf=Table[Flatten[{freq[j,1],Qmim[j]}],{j,1,iodd}];
Print[Qgraf];

```

END

28|05|92

FIN