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# MM-WAVE INTEGRATED RF-MEMS TUNABLE CAVITY RESONATORS, FILTERS AND ULTRA-LOW PHASE-NOISE OSCILLATORS

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*To my parents.*



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Finally, I want to dedicate this text to my parents. This is to make sure I thank them for everything they had done so far for me and for their love. Their unconditional support made it possible.

*Dragos Dancila*  
*Louvain-la-Neuve, June 2011*



# ABSTRACT

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Within the Microelectronics Industry, an essential part of development is dedicated to the fabrication of high-volume and low-cost transistor based electronics. These devices are essential for the mass consumer electronics markets. The main research effort is focused on the realization of the Moore's law, which states that circuit densities double every 24 months, shaping the framework of the *More of Moore* paradigm. This means that the price per devices is reduced by a factor two, within this time frame. However, Moore's law is expected to end, as devices are reaching limitations inherent to the approach of the atomical dimensions. Alternative research paths emerged forming the new *More than Moore* paradigm. This consists in using the microfabrication technological know-how towards the realization of alternative devices and applications: energy scavengers, monitoring medical devices, chemical sensors... and among others the miniaturization of radio devices. Among these research topics, this PhD focuses on the integration of Radio Frequency (RF) - Micro Electromechanical Systems (MEMS), i.e. RF-MEMS devices. The RF-MEMS devices already compete against electronic components e.g. capacitors, inductors, resonators, varicaps, transmission lines, etc... as they offer high performance, tuning by movable parts and open new perspectives at extra high frequency i.e. 30 to 300 GHz. Subsequently, the research on RF-MEMS cavity resonators stems since the late nineties. The first realizations show comparable performance, in terms of quality factors to the metallic waveguides. The latest are quite bulky, high in weight, large in size, costly to manufacture and difficult to integrate with electronic circuits. The RF-MEMS cavity resonators alleviate these inconveniences, while they are fully integrated. However, voltage controlled integrated tuning systems were missing at extra high frequencies.

In this thesis, cavity resonators' design and characterization are introduced, as a preliminary discussion. The coupling mechanism is thoroughly investigated as well as different feeds. The integration in bulk silicon by micromachining techniques is realized at 60 and 75 GHz. Further, the tuning performance induced by internal volumes of perturbation is investigated. Furthermore, the integration of a voltage controlled tuning system for air cavity resonator is realized at 60 GHz, using a MEMS based Faraday cage. Additionally, a new miniaturization concept is demonstrated using high impedance surfaces. As a passive application using cavity resonators, a seven-pole Chebyshev bandpass filter is realized in Low Temperature Co-fired Ceramic technology. The filter is characterized by a large bandwidth of 7 GHz, covering the full unlicensed spectrum around 60 GHz and presents sharp stopband attenuation. Finally, an ultra-low phase-noise oscillator at 60 GHz is realized as an active application using substrate integrated cavity resonators to improve its performance. Actually, the demand for mm-wave ultra-low phase-noise oscillators is supported by cars manufacturers, massively introducing automotive RADARs for collision prevention. This

oscillator improves the State of the Art for integrated oscillators in the frequency band from 40 to 80 GHz by demonstrating the highest factor of merit, to our best knowledge and to date,  $\text{FoM} = -199 \text{ dBc/Hz @ } 1 \text{ MHz offset from the carrier frequency}$ ,  $f_{osc} = 59.98 \text{ GHz}$ . The latest is due to the combination of both an ultra-low phase-noise of  $-116 \text{ dBc/Hz @ } 1 \text{ MHz offset from the carrier frequency}$  and relatively high output power,  $P_{OUT} = 10.03 \text{ dBm}$ .



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# GLOSSARY

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## Acronyms and Abbreviations

|               |                                                      |
|---------------|------------------------------------------------------|
| < 100 >       | Silicon crystal orientation                          |
| III-V         | Composite semiconductor from the III and V groups    |
| 2D            | Two Dimensional                                      |
| 3D            | Three Dimensional                                    |
| 3D $\mu$ Tune | European Project (IST-2005-027768)                   |
| AAC           | Adaptive Cruise Control                              |
| ADS           | Advanced Design System, RF simulation Software       |
| AMC           | Artificial Magnetic Conductors                       |
| BAW           | Bulk Acoustic Wave resonators                        |
| Bi-CMOS       | Bipolar Complementary Metal-Oxide-Semiconductor      |
| BW            | Bandwidth                                            |
| CB            | Common Base                                          |
| CE            | Common Emitter                                       |
| CG            | Common Gate                                          |
| CMOS          | Complementary Metal-Oxide Semiconductor              |
| CPP           | Central Post Perturbation                            |
| CPW           | Co-Planar Waveguide                                  |
| CS            | Common Source                                        |
| C-V           | Capacitance versus Voltage                           |
| dB            | Decibel                                              |
| dBc           | Decibel relative to the carrier                      |
| dBm           | Decibel with the reference quantity of one milliwatt |
| DR            | Dielectric Resonator                                 |
| DRIE          | Deep Reactive-Ion Etching                            |
| EFFA          | Electrostatic Fringing Field Actuator                |
| FBAR          | Film Bulk Acoustic Resonator                         |
| FBW           | Fractional bandwidth                                 |

|          |                                                                                    |
|----------|------------------------------------------------------------------------------------|
| FC       | Flip-Chip                                                                          |
| FET      | Field-Effect Transistor                                                            |
| FFA      | Feed From Above                                                                    |
| FFB      | Feed From Below                                                                    |
| FoM      | Factor of Merit                                                                    |
| G        | Closed loop gain                                                                   |
| GHz      | Giga Hertz ( $10^9$ Hz)                                                            |
| Gunn     | Gunn effect diode                                                                  |
| HB       | Harmonic Balance                                                                   |
| HBT      | Heterojunction Bipolar Transistor                                                  |
| HEMT     | High Electron Mobility Transistor                                                  |
| HF       | High Frequency                                                                     |
| HFSS     | High Frequency Structural Simulator                                                |
| HIS      | High Impedance Surfaces                                                            |
| Hz       | Hertz                                                                              |
| ICTEAM   | Information and Communication Technologies,<br>Electronics and Applied Mathematics |
| IEEE     | Institute of Electrical and Electronics Engineers                                  |
| IL       | Insertion Loss                                                                     |
| IMEC     | Interuniversity Microelectronics Center                                            |
| IMPATT   | IMPact ionization Avalanche Transit-Time diode                                     |
| Ka-band  | IEEE frequency band 26.5 - 40 GHz                                                  |
| KHz      | Kilo Hertz ( $10^3$ Hz)                                                            |
| KOH      | Heated Potassium Hydroxide                                                         |
| KULeuven | Katholieke Universiteit Leuven                                                     |
| LIGA     | Lithographie Galvanoforming Abformung                                              |
| LNA      | Low Noise Amplifier                                                                |
| LPCVD    | Low-Pressure Chemical Vapor Deposition                                             |
| LTCC     | Low Temperature Co-fired Ceramic                                                   |
| MCM-D    | Multi-Chip Module-Deposited                                                        |
| Meas.    | Measurements                                                                       |
| MEMS     | Micro Electro Mechanical System                                                    |
| MESFET   | Metal semiconductor Field Effect Transistor                                        |
| MHz      | Mega Hertz ( $10^6$ Hz)                                                            |
| MIM      | Metal-Insulator-Metal                                                              |

|          |                                                      |
|----------|------------------------------------------------------|
| M-M      | Male-Male                                            |
| mm-wave  | Millimeter wave                                      |
| MOSFET   | Metal-Oxide-Semiconductor Field-Effect Transistor    |
| PCB      | Printed Circuit Board                                |
| PEC      | Perfect Electric Conductor                           |
| PECVD    | Plasma Enhanced Chemical Vapor deposition            |
| PhD      | Doctor of Philosophy                                 |
| piezo    | Piezoelectric                                        |
| PLL      | Phase-locked loop                                    |
| PMC      | Perfect Magnetic Conductor                           |
| PMMA     | Poly(methylmethacrylate)                             |
| PNA-X    | Microwave Network Analyzer                           |
| ppm      | parts-per-million ( $10^{-6}$ )                      |
| PVD      | Physical vapor deposition                            |
| Q-factor | Quality factor                                       |
| QZERO    | Kajfez's software [ <a href="#">Kajfez 94</a> ]      |
| RADAR    | RAdio Detection And Ranging                          |
| RBW      | Resolution Band Width                                |
| RF       | Radio Frequency                                      |
| RF-MEMS  | Radio Frequency Micro Electro Mechanical System      |
| RIE      | Reactive Ion Etching                                 |
| RL       | Return Loss                                          |
| SAP      | Side Annular Perturbation                            |
| SAW      | Surface Acoustic Wave resonators                     |
| SEM      | Scanning Electron Microscopy                         |
| SiCR     | Silicon integrated Cavity Resonator                  |
| SiCRO    | Silicon integrated Cavity Resonator based Oscillator |
| SiGe     | Silicon Germanium                                    |
| Sim.     | Simulations                                          |
| SIW      | Substrate Integrated Waveguide                       |
| SOI      | Silicon-on-Insulator                                 |
| SOLT     | Short-Open-Load-Thru, calibration method             |
| ss       | Small signal                                         |
| TE       | Transverse Electric                                  |
| TEM      | Transverse Electro Magnetic                          |

|        |                                          |
|--------|------------------------------------------|
| T-line | Transmission line                        |
| TLR    | Thru-Line-Reflection, calibration method |
| TM     | Transverse Magnetic                      |
| TSM    | Top Surface Metallization                |
| TSV    | Through-Substrate Vias                   |
| TTV    | Total Thickness Variation                |
| UCL    | Université catholique de Louvain         |
| V-band | IEEE frequency band 50 - 75 GHz          |
| VBW    | Video Band Width                         |
| VCO    | Voltage-Controlled Oscillator            |
| VNA    | Vector Network Analyzer                  |
| VTT    | Technical Research Center of Finland     |
| WG     | Waveguide                                |
| WLAN   | Wireless Local Area Networks             |
| WR15   | V-Band waveguide                         |
| X-band | IEEE frequency band 7 - 12 GHz           |
| XRL    | X-Ray Lithography                        |

### List of Symbols

|                                |                                                               |
|--------------------------------|---------------------------------------------------------------|
| $a, d, b$                      | Width, length and height of a cavity resonator                |
| $a_1, a_2, a_3$                | Transformation coefficients for data fitting                  |
| $A, \alpha$                    | Parameters resulting the extraction of $f_0$                  |
| $a_{KOH}, d_{KOH}$             | Width, length of the KOH-etched cavity                        |
| $\alpha_1, \alpha_2, \alpha_3$ | Parameters resulting the extraction of the equivalent circuit |
| $A_{mn}$                       | Amplitudes of the $m, n^{th}$ resonance mode                  |
| $A_p$                          | Ripples attenuation                                           |
| $B$                            | Susceptance                                                   |
| $\beta$                        | Coupling coefficient at resonance ( $\kappa(f_0)$ )           |
| $\beta_{mn}$                   | Propagation constant of the $m, n^{th}$ mode                  |
| $B_i$                          | Shunt susceptance of coupling iris                            |
| $bps/Hz$                       | Spectral efficiency measured in bit per second per Hz         |
| $C$                            | Capacitor                                                     |
| $c_0$                          | Speed of light (299 792 458 m/s)                              |
| $C_{end}$                      | Capacitance of the MEMS after the actuation cycle             |



|                    |                                                                      |
|--------------------|----------------------------------------------------------------------|
| $CPW_s$            | Slot of the CPW line                                                 |
| $CPW_w$            | Width of the CPW line                                                |
| $C_{start}$        | Capacitance of the MEMS before the actuation cycle                   |
| $D$                | Diameter of the SiCR                                                 |
| $D_1$              | Diameter of the resonance locus, in the lossless case                |
| $D_2$              | Auxiliary diameter in the case of coupling losses                    |
| $D_{eff}$          | Effective diameter                                                   |
| $\delta$           | Normalized frequency ratio                                           |
| $\delta_{EFA}$     | Measured initial deflection of the EFA                               |
| $\delta_s$         | Skin depth $\left(\sqrt{\frac{2}{\omega\mu\sigma}}\right)$           |
| $\Delta V$         | Volume of the perturbation box in the cavity resonator               |
| $d_i$              | Opening length of coupling iris                                      |
| $d_x, d_y$         | Cross section of the feeding current $I_0$ , located at $(x_0, y_0)$ |
| $\vec{E}$          | Electric field vector                                                |
| $E_{AlCu_{0.5\%}}$ | Strain of the AlCu <sub>0.5%</sub> alloy                             |
| $e_{mn}$           | Normalized $E_z$ fields of the $m, n^{th}$ resonance mode            |
| $\eta$             | Wave impedance $\left(\frac{\eta_0}{\sqrt{\epsilon_r}}\right)$       |
| $\eta_0$           | Wave impedance of free space (377 $\Omega$ )                         |
| $\eta_{eff}$       | Effective shielding coefficient                                      |
| $F$                | Active device noise figure at operating power                        |
| $f_{\pm Q_0}$      | Intersecting resonance frequencies with the loci of $Q_0$            |
| $f_{\pm Q_e}$      | Intersecting resonance frequencies with the loci of $Q_e$            |
| $f_{\pm Q_l}$      | Intersecting resonance frequencies with the loci of $Q_l$            |
| $f_{0,HIS}$        | Resonance frequency of the HIS metallic patch                        |
| $f_{even}$         | Resonance frequency of the cavity resonator, PEC iris                |
| $f_{LO}$           | Local Oscillator frequency                                           |
| $f_{mnl}$          | Resonance frequency of the $m, n, l^{th}$ mode                       |
| $f_{odd}$          | Resonance frequency of the cavity resonator, PMC iris                |
| $f_{osc}$          | Oscillation frequency                                                |
| $f_0$              | Unloaded resonance frequency                                         |
| $f_c$              | Central frequency                                                    |
| $f_m$              | Offset frequency at which the phase noise is evaluated               |
| $f_r$              | Loaded resonance frequency                                           |

|                    |                                                                                 |
|--------------------|---------------------------------------------------------------------------------|
| $FF_{TM010}$       | Form factor of the $TM_{010}$ cylindrical resonator                             |
| $f_{off}$          | Cutoff frequency                                                                |
| $\Gamma$           | Reflection coefficient                                                          |
| $\Gamma_{EFFA}$    | Strain gradient of the EFFA                                                     |
| $\Gamma_A$         | Reflection coefficient of the active circuit                                    |
| $\Gamma_d$         | De-tuned reflection coefficient                                                 |
| $\Gamma_i$         | Input reflection coefficient                                                    |
| $\Gamma_R$         | Reflection coefficient of the resonator                                         |
| $\Gamma_r$         | The reflection coefficient which is the closest to the origin                   |
| $g_i$              | Chebyshev prototype's element value                                             |
| $G_{mn}$           | Normalized form factor                                                          |
| $\vec{H}$          | Magnetic field vector                                                           |
| $H_{box}$          | Height of the perturbation rectangular post                                     |
| $H_{cage}$         | Height of the Faraday cage, similar to $H_{box}$                                |
| $H_{HR-Si}$        | Height of the HR-Si                                                             |
| $h_{PCB}$          | Thickness of the PCB substrate                                                  |
| $H_t$              | Tangential magnetic field                                                       |
| $I$                | DC Current                                                                      |
| $I_{b_{RF}}$       | RF base current                                                                 |
| $I_d$              | Drain current                                                                   |
| IF                 | Intermediate Frequency                                                          |
| $\vec{J}$          | Current distribution                                                            |
| $k_0$              | Propagation constant of free space ( $\omega\sqrt{\mu_0\varepsilon_0}$ )        |
| $\kappa$           | Coupling coefficient                                                            |
| $\kappa_m$         | Magnetic coupling                                                               |
| $\kappa_{norm.}$   | Normalized coupling coefficient using the ratio: $\frac{Q_{0Meas.}}{Q_{0Sim.}}$ |
| $k_{i,j}$          | Filter's internal coupling coefficient                                          |
| $k_{mnl}$          | Wavenumber of the $m, n, l^{th}$ mode                                           |
| $kT$               | Thermal noise (-174 dBm at room temperature)                                    |
| $L$                | Inductor                                                                        |
| $\lambda$          | Wavelength                                                                      |
| $\lambda_g$        | Waveguide wavelength                                                            |
| $L_{EFFA}$         | Length of the EFFA                                                              |
| $\mathcal{L}(f_m)$ | Phase noise evaluate at $f_m$ offset from the carrier frequency                 |

|              |                                                            |
|--------------|------------------------------------------------------------|
| $l_i$        | Physical length of the waveguide between two irises        |
| $\mu$        | Permeability                                               |
| $\mu m$      | Micrometer                                                 |
| $\mu_0$      | Permeability of free space ( $4\pi \times 10^{-7} H/m$ )   |
| $\mu_r$      | Relative permeability                                      |
| $N$          | Number of poles                                            |
| $n$          | Transformer coefficient                                    |
| $o$          | Diameter of the circular aperture                          |
| $\omega$     | Angular frequency ( $2\pi f$ )                             |
| $\omega_0$   | Resonance angular frequency of a LC-tank                   |
| $p$          | Diameter of the metallic patch feed                        |
| $P_0$        | Power dissipated in the resonator                          |
| $P_{cond}$   | Power dissipated in conductive losses                      |
| $P_c^p$      | Total power losses for a cavity with a perturbation        |
| $P_d$        | Average power dissipated                                   |
| $P_{diel}$   | Power dissipated in the lossy dielectric                   |
| $P_e$        | Power dissipated in the external circuit                   |
| $\phi_i$     | Electrical length of the waveguide between two irises      |
| $p_l, p_w$   | CPW additional aperture's length and width                 |
| $P_{out}$    | Output power                                               |
| $P_{rad}$    | Power dissipated in radiative losses                       |
| $Q_0$        | Unloaded Q-factor                                          |
| $Q_{0Meas.}$ | Unloaded Q-factor extracted from measurements              |
| $Q_{0Sim.}$  | Unloaded Q-factor extracted from HFSS simulations          |
| $Q_{cond}$   | Unloaded Q-factor, only conductive losses are considered   |
| $Q_{diel}$   | Unloaded Q-factor, only dielectric losses are considered   |
| $Q_e$        | External Q-factor                                          |
| $Q_l$        | Loaded Q-factor                                            |
| $Q_{rad}$    | Unloaded Q-factor, only radiation losses are considered    |
| $R$          | Resistor                                                   |
| ratio surf.  | Ratio between the aperture's surface and the cavity's area |
| $R_c$        | Characteristic impedance                                   |
| $R_s$        | Surface resistance                                         |
| $S_{11}$     | Reflection coefficient                                     |
| $S_{11A}$    | Active device reflection coefficient                       |

|                    |                                                                |
|--------------------|----------------------------------------------------------------|
| $S_{11R}$          | Resonator reflection coefficient                               |
| $\sigma$           | Material conductivity                                          |
| slot               | Slot of the HIS metallic patch                                 |
| $t$                | Fractional transformation of the normalized frequency          |
| $T$                | Period of oscillation                                          |
| $\tan \delta$      | Loss tangent                                                   |
| $V$                | DC Voltage                                                     |
| $\varepsilon$      | Permittivity                                                   |
| $\varepsilon_0$    | Permittivity of free space ( $8.854184310^{-12} \text{ F/m}$ ) |
| $\varepsilon_r$    | Relative permittivity                                          |
| $V_{be}$           | Base Emitter Voltage                                           |
| $V_{ce}$           | Collector Emitter Voltage                                      |
| $V_{mc}$           | Effective volume of the coupling aperture                      |
| $V_o$              | Volume of the cavity resonator                                 |
| $V_{PI}$           | Pull-in voltage                                                |
| $V_{PO}$           | Pull-out voltage                                               |
| $w$                | Width of the ML feed's connecting microstrip line              |
| $W, L$             | Width, length of the perturbation rectangular post             |
| $w_0, l_0$         | Width and length of the LTCC cavity resonator                  |
| $W_d$              | Dissipated energy in the resonator during one period           |
| $W_{e,m}$          | Electric, Magnetic energy stored in the cavity resonator       |
| $W_{e,m_{CPP}}$    | Electric, Magnetic energy stored in the CPP                    |
| $W_{e,m_{SAP}}$    | Electric, Magnetic energy stored in the SAP                    |
| $W_{e,m_{UP}}$     | Electric, Magnetic energy stored in the volume UP              |
| $W_{HIS}, L_{HIS}$ | Width, Length of the HIS metallic patch                        |
| $W_{max}$          | Maximum energy stored in the resonator                         |
| $X$                | Reactance                                                      |
| $x_{WG}$           | Normalized reactance of an aperture in a waveguide             |
| $\xi$              | Normalized frequency difference                                |
| $Y$                | Admittance                                                     |
| $y$                | Normalized admittance                                          |
| $Y_0$              | Admittance of the lossless LC-tank                             |
| $Z$                | Impedance                                                      |
| $z$                | Normalized impedance                                           |
| $Z_A$              | Impedance of the active port                                   |

|          |                                                       |
|----------|-------------------------------------------------------|
| $Z_c$    | Characteristic impedance (50 $\Omega$ )               |
| $Z_R$    | Impedance of the resonator port                       |
| $Z_0$    | Impedance of the lossless LC-tank                     |
| $Z_{in}$ | Impedance of the input terminal of a cavity resonator |
| $Z_{TE}$ | Wave impedance of TE propagation mode                 |

**Materials**

|                                |                          |
|--------------------------------|--------------------------|
| Ag                             | Silver                   |
| AlCu <sub>0.5%</sub>           | Copper Aluminum alloy    |
| AlN                            | Aluminum Nitride         |
| Au                             | Gold                     |
| BCB                            | Benzo-CycloButene        |
| Cu                             | Copper                   |
| GaAS                           | Galium Arsenide          |
| HR-Si                          | High Resistivity Silicon |
| InP                            | Indium Phosphide         |
| LR-Si                          | Low Resistivity Silicon  |
| Ni                             | Nickel                   |
| Si                             | Silicon                  |
| SiN                            | Silicon Nitride          |
| Sn                             | Tin                      |
| Ta <sub>2</sub> O <sub>5</sub> | Tantalum Pentoxide       |
| Ti                             | Titanium                 |



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*Success is never found. Failure is never fatal.  
Courage is the only thing.*

***Sir Winston Churchill***





## CHAPTER 1

---

## INTRODUCTION

---

The Microelectronics Industry is largely dominated by "Moore's law", which is governing its innovation cycles. This law states that circuit densities double every 24 months, based on Gordon Moore's observation, five decades ago [[Moore 98](#)].

Initial very low yields of device manufacturers turned the semiconductor industry's development towards the improvement of the yield by the standardization of the production methods. As a result, a viable semiconductor industry emerged but in turn "the technology led the science in a sort of inverse linear model" [[Schaller 96](#)], as the research was focused principally on development. The trend continues today within the framework of the *More of Moore* approach. This approach constitutes the core business of the semiconductor industry. Moore's law still holds, drifting into an economical requirement from a technological observation. The fulfillment of such a requirement is securing industry's players to sustain their growth.

*"Moore's Law is a sort of technological barometer. It very clearly tells you that if you take the information processing power you have today and multiply by two, that will be what your competition will be doing 18 months from now. And that is where you too will have to be."* [[Malone 96](#)]

Besides this, business history is full of examples of missed opportunities, the so called "Marketing Myopia" phenomena [[Levitt 60](#)]. Actually, none of the major semiconductor players today were dominant electronics firms at the beginnings. Furthermore, Moore's law is expected to stop around 2015 because devices, with critical dimensions below 10 nanometers are reaching limitations inherent to the approach of the atomical dimensions.

Therefore, alternative research paths were encouraged and emerged from the semiconductor industry's core. Emerging research flourished, stemming from the available technological knowhow. However, research implementations are still too diverse to be focused and optimized at the same level as the core of the semiconductor industry. Nevertheless, the ground is open for a large range exploration, forming the so called *More than Moore* path. Among the endogenous research topics, we point out energy scavengers, nano-devices as plasmonics and spintronics, medical devices as neuro-probes, fluidics, ... among others the miniaturization of integrated Radio Frequency (RF) and Micro Electromechanical Systems (MEMS) devices, triggering our interest in this thesis.

The term **RF** refers to a frequency range for electromagnetic waves, typically from 300 MHz ( $3 \times 10^8$  Hz) to 300 GHz ( $3 \times 10^{11}$  Hz) with a corresponding electrical wavelength between  $\lambda = c_0/f = 1$  m and  $\lambda = 1$  mm, respectively [Pozar 05]. More specifically, signals with wavelengths on the order of the millimeter are called *millimeter waves* (mm-wave).

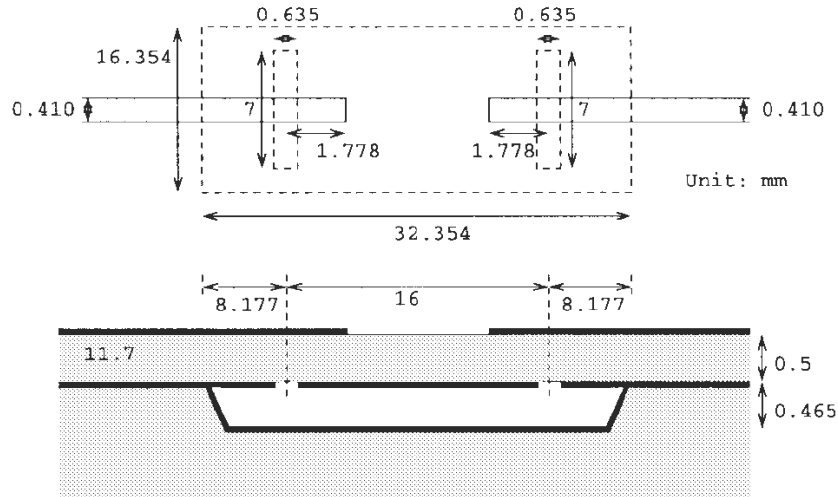
The **miniaturization** and **integration** are driving forces of the wireless communication systems. Since, back to the 12<sup>th</sup> of December 1901, when Marconi used for the first time a 48 m monopole antenna stretched between two masts distant of 60 m, for signal transmission across the Atlantic [Ramsay 81], the range of frequencies and dimensions were continuously reduced to the actual RF technology investigations in the mm-wave domain. New standards for wireless communication consumer applications, corresponding to frequencies from 30 to 100 GHz appeared during the last years. Prominent examples of this development are especially automotive RADARs at 77 GHz, satellite communication in different frequency intervals up to 64 GHz and very recently also, 60 GHz high data rate Wireless Local Area Networks (WLAN). The challenge is to find a mix of technologies and components that lead to simpler, highly integrated, cheap and reliable systems. Indeed, integrated passives as inductors for low cost (Bi)CMOS or Bipolar technologies, used for application below 10 GHz, are typically characterized by a low **quality factor (Q-factor)**, limited to around 10 and with the expense of higher processing complexity and cost, up to 20-30 [Tilmans 03]. Additionally, the higher the operation frequency the lower the performance. Instead, cavity resonators made by entirely enclosed metallic rectangular or cylindrical waveguides achieve very high unloaded Q-factors, as high as thousands, also at very high frequencies [Pozar 05]. However, conventional waveguide cavity resonators are heavy in weight, large in size and costly to fabricate. Meanwhile, these waveguides can be realized in an integrated way, with similar performance, using micromachining, novel and alternative techniques.

The **MEMS** term is used to define both the technology and the devices realized in these technologies. The MEMS are made of components between 1 to 100  $\mu\text{m}$  in size and MEMS devices generally range in size from 20  $\mu\text{m}$  to a millimeter. The technology enables the manufacturing of mechanical structures *on* and *in wafer* [MEMS]. *In wafer* technologies are using the whole thickness of a silicon wafer for building micro-mechanical structures. Silicon is machined using various etching processes as KOH, wet etch or Deep Reactive-Ion Etching (DRIE), dry etch. Sputtering and plating techniques are used for adding features in the third dimension and could also make possible to achieve hermetic encapsulation. *On wafer* technologies correspond to sur-

face micromachining using successive layers deposition on the surface of a substrate as the structural materials, rather than using the substrate itself.

## 1.1 CONTEXT OF THE THESIS

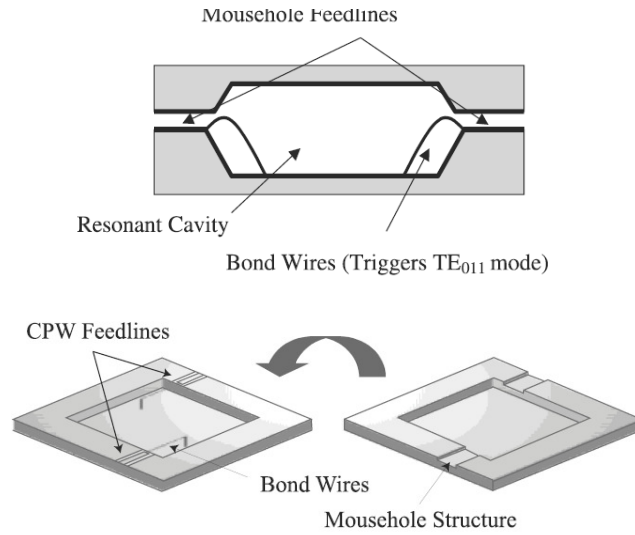
The research on micromachined cavity resonators stems since the late nineties, flourishing with the development of micromachining techniques. These techniques allow the realization of high precision cavity resonators, by highly anisotropic etch processes (KOH, DRIE) in low (LR-Si) or high resistivity silicon (HR-Si), followed by diverse metallizations. At high frequencies (above 10 GHz) typical sizes are small and compatible with the micromachining techniques. Here follows the highlights of the State of the Art of fixed in frequency and tunable integrated cavity resonators, summarized chronologically in Table 1.1. The first micromachined cavity resonator realization is shown in Figure 1.1. Comparable performance to the metallic waveguide (WG) is demonstrated. Additionally, these are integrated on-chip [Papapolymerou 97, Harle 03].



Papapolymerou et al., “A micromachined high-Q X-band resonator,”  
*IEEE Microwave and Guided Wave Letters*, June, 1997.

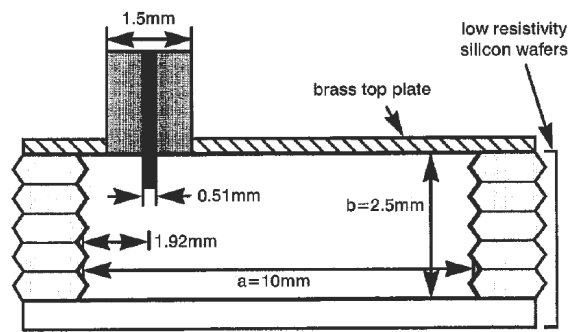
Fig. 1.1.: Micromachined high-Q (506) integrated in low resistivity silicon (LR-Si) air cavity resonator at the resonance frequency,  $f_0 = 10.285$  GHz [Papapolymerou 97].

The Q-factor is limited by the height of the cavity resonator, which itself is limited to the standard thickness of the wafer. To surmount this, Brown *et al.* proposed a cavity resonator formed by two standard ( $525 \mu\text{m}$  thick) silicon wafers, etched to  $450 \mu\text{m}$  deep and assembled together, as can be seen in Figure 1.2. A CPW line, prolonged by a wire bonded to the base of the cavity is ensuring the feeding of the  $\text{TE}_{011}$  mode of resonance. The measured Q-factor is 1117 at 23.97 GHz.



Brown et al., "Microwave and millimeter-wave high-Q micromachined resonators,"  
*International Journal of RF and Microwave Computer-Aided Engineering*, July, 1999.

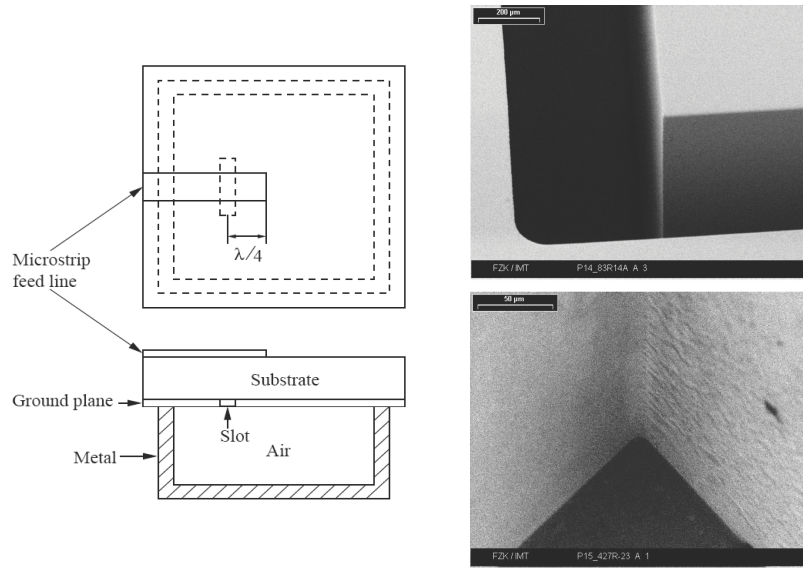
Fig. 1.2.: Double wafer thick millimeter-wave high-Q (1117) micromachined air cavity resonator ( $f_0 = 23.97$  GHz), in LR-Si [Brown 99].



Stickel et al., "High-Q bulk micromachined silicon cavity resonator at Ka-band,"  
*Electronics Letters*, March, 2001.

Fig. 1.3.: High-Q (2155-4520) micromachined silicon (stacking of five wafers) air cavity resonator at Ka-band ( $f_0 = 30$  GHz) [Stickel 01].

As deeper cavity resonators are characterized by higher Q-factors, Stickel *et al.* demonstrated a high-Q integrated air cavity resonator, composed of several stacked standard silicon wafers, fully etched through their thickness. In [Stickel 01], shown in Figure 1.3, a Q-factor of 2155 at 30 GHz, for a total thickness of 2.5 mm (stacking of five wafers) is demonstrated. In [Stickel 03], a Q-factor of 4520 at 30 GHz is demonstrated for a total thickness of 7.5 mm (stacking of two times six wafers). However the performance is excellent, the stacking is challenging as air gaps are deteriorating the Q-factor. It was also demonstrated that the wafers can be joined by the electroplating of a relatively thick layer (100  $\mu\text{m}$ ) of copper metallization. This further eliminates any residual gaps between wafers. Additionally, several issues were reported, concerning the connectivity of the  $\text{TE}_{101}$  cavity resonator, at its top and bottom, see Figure 1.3. However, high performance was obtained by joining two cavities such that, for the propagating mode, there is no current flowing across the melding plane. Considering the surface currents of the  $\text{TE}_{101}$  mode, emerging from the center top and bottom sides, the mating plane is ideally located in the center of the cavity resonator, parallel to the E field and perpendicular to the propagation plane of the equivalent  $\text{TE}_{10}$  waveguide [Stickel 03].

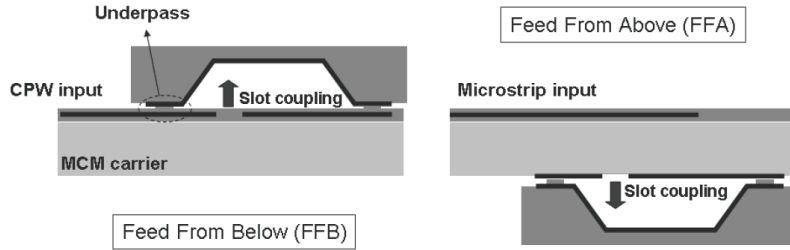


Zhen Ma, "GHz LIGA cavity resonators and filters for microwave and millimetre-wave applications," *University of Saskatchewan, PhD, 2007*.

Fig. 1.4.: LIGA high-Q (2122) air cavity resonator ( $f_0 = 24$  GHz). Insert of SEM pictures of the patterned PMMA resist before electro-plating [Ma 07].

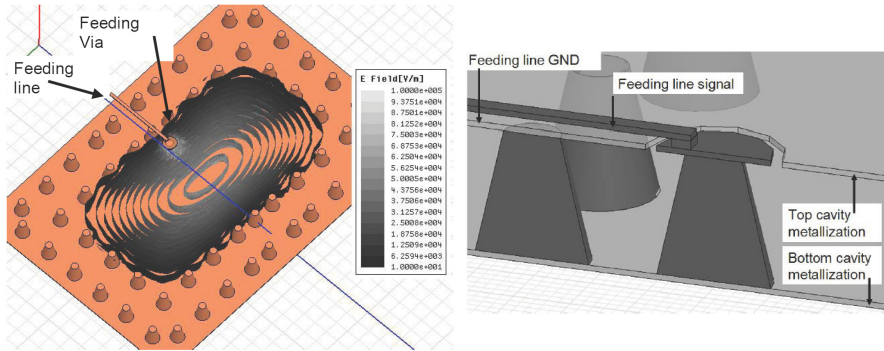
Alternatively, Ma *et al.* investigates deep X-ray Lithography (XRL) of the LIGA process as a microfabrication technology for structuring polymers (i.e. PMMA), towards the realization of RF/microwave resonators and filters. The term LIGA is an acronym for the German Lithography, Electroplating and Moulding. This technology offers sub-micron features, aspect ratios of 100:1 or higher for resist thicknesses of up to 3 mm,

and almost vertical and optically smooth sidewalls [Ma 07]. An air cavity resonator with a Q-factor of 2122 at 24 GHz, for a total thickness of 2 mm is demonstrated, see Figure 1.4.



Ilja Ocket, “Millimeter Wave Antennas, Circuits and Systems in MCM-D Technology,”  
Katholieke Universiteit Leuven, PhD, 2009.

Fig. 1.5.: Micromachined cavity resonators assembled on top and at the bottom of IMEC’s MCM-D passive integration platform. High Q-factors 744 and 584 at respectively 60 and 74.16 GHz are demonstrated [Ocket 09, Ocket 10].



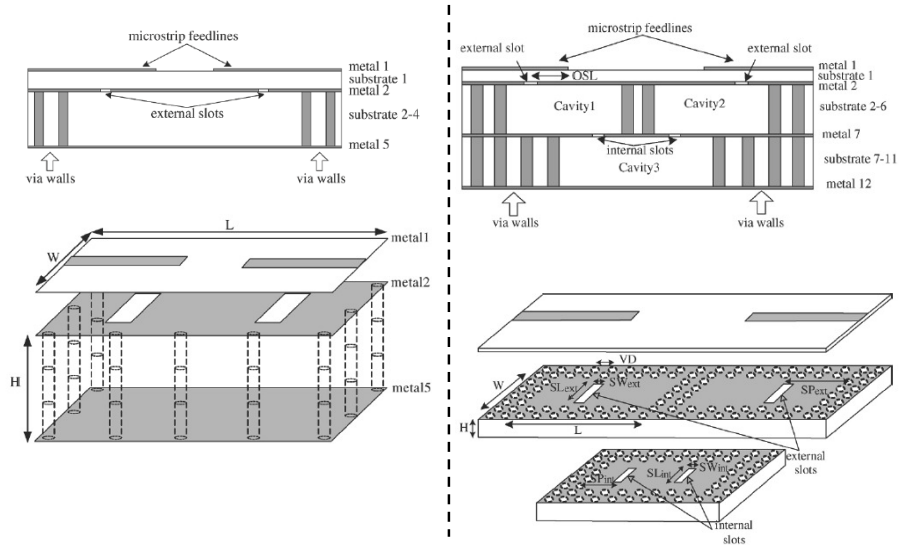
Posada et al., “Ultra-Miniaturized Integrated Cavities on High-Resistivity Silicon Thin-Film MCM-D Technology,” *IEEE Topical Meeting on Silicon*

Fig. 1.6.: MCM-D substrate integrated cavity resonators demonstrated Q-factor of 170 and 186 at respectively 29 and 60 GHz [Posada 08].

At higher frequencies, Ocket *et al.* reports on 60 and 74 GHz cavity resonators realized as an assembly of KOH-etched micromachined cavities on top of IMEC’s passive integration platform, MCM-D. The assembly can be done in two ways, as can be seen in Figure 1.5. A first option, called Feed From Below (FFB) consists in assembling the cavities on the top side of the MCM-D carrier. This is the least complicated ap-

proach but requires the interconnect line to pass under or through the bonding ring to couple the cavity and requires some standoff between the cavity and other mounted components. In this approach the cavity can be coupled by a CPW antenna feed in the cavity bottom plane. In a second approach, called Feed From Above (FFA) the cavity is assembled on the backside of the carrier and is connected with a microstrip line in one of the top Cu layers. Its ground plane is the backside metal. High Q-factors 744 and 584 at 60 and 74.16 GHz respectively are demonstrated [Ocket 09, Ocket 10].

In the same technology, Posada *et al.* integrated cavity resonators in the bulk high resistivity silicon (HR-Si), using available Through-Substrate Vias (TSV), punching through the  $100\ \mu\text{m}$  thick wafer. The cavities are shaped by a double or a single fence of TSVs, with a via diameter of  $100\ \mu\text{m}$  and a pitch of  $220\ \mu\text{m}$ . He demonstrated that one row of vias is enough to eliminate radiation at 29 GHz, but at 60 GHz two rows vias are necessary. These integrated cavity resonators have the potential to provide very good performance (Q-factor of 170 and 186 at respectively 29 and 60 GHz) while they are small in size (3.4 times smaller than air cavities) and *de-facto* fully integrated [Posada 08].



Lee et al., "Low-loss LTCC cavity filters using system-on-package technology at 60 GHz," *IEEE Transactions on Microwave Theory and Techniques*, December, 2005.

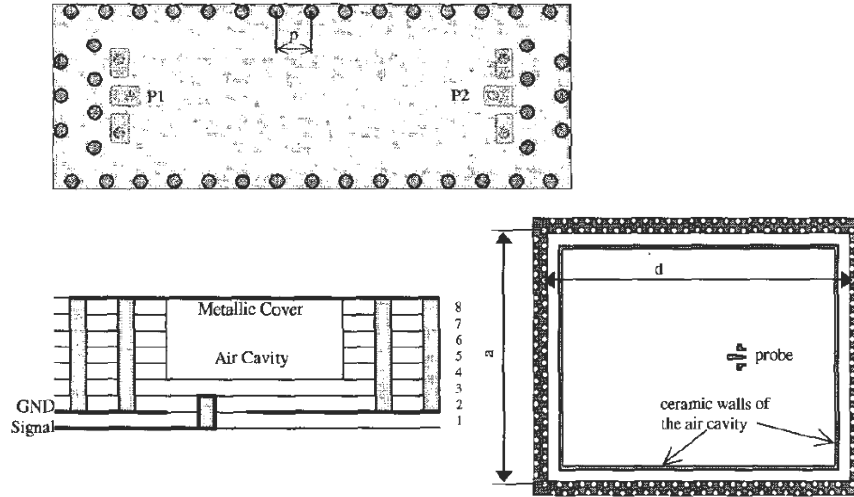
Fig. 1.7.: SIW cavity resonator and filter, integrated in LTCC demonstrated a Q-factor of 367 at 60 GHz [Lee 05a].

Similarly, Substrate Integrated Waveguides (SIW) defined by rows of vias shaping cavity resonators and filters are implemented in Low Temperature Co-fired Ceramic (LTCC) technologies [Lee 05a, Chen 07, VTT]. The performance of these resonators is

limited by the dielectric loss, although the material has a low loss tangent. A resonator with a Q-factor of 367, for a total thickness of 0.3 mm (stacking of three LTCC layers) is demonstrated in [Lee 05a], as can be seen in Figure 1.7. A three-pole bandpass filter at 60 GHz, with an Insertion Loss (IL) of 2.14 dB at the center frequency of 58.7 GHz, a rejection loss of 16.4 dB over the passband, and a 3-dB bandwidth approximately 1.38% (0.9 GHz) are also presented.

To avoid dielectric losses in LTCC, El-Tager proposed to remove the dielectric while shaping the cavity resonator by two rows of vias embedded in ceramic, as can be seen in Figure 1.8. A resonator with a Q-factor of 1206, for a total thickness of 0.7 mm (stacking of seven LTCC layers) is demonstrated at 19.8 GHz.

In another LTCC realization, the miniaturization of cavity resonators is realized by incorporating a capacitive post in the middle of a cavity resonator, reducing its lateral dimensions to  $1.45 \times 1.45 \times 0.4 \text{ mm}^3$ , at 42.6 GHz. The resonator is characterized by a 52% size reduction at the expense of a Q-factor reduction by 20%, compared to the unloaded cavity at the same frequency [Ferrand 05].



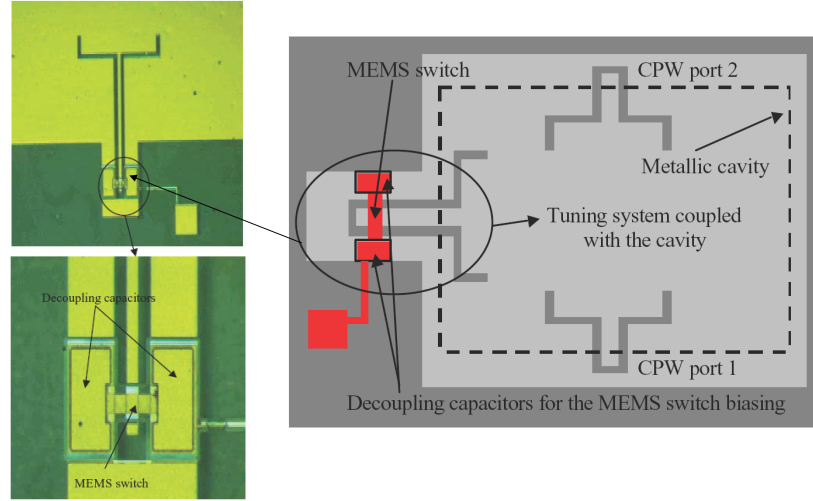
El-Tager et al., “High-Q LTCC resonators for millimeter wave applications,”  
*IEEE MTT-S International Microwave Symposium Digest*, June, 2003.

Fig. 1.8.: Air cavity resonator shaped in LTCC improving the Q-factor (1206) at 19.8 GHz over standard LTCC designs [El-Tager 03].

Besides fixed frequency cavity resonators, there is a need for tunable cavity resonators for use in tunable filters and oscillators [Tilmans 07]. Several Universities, mainly in the United States, anticipated this need with publications on tunable integrated, micromachined high-Q cavity resonators. Actually, quite few examples of integrated tunable cavity resonators are available, as can be seen in Table 1.1, among which



two trends can be identified. In the first approach, a lumped varactor is loading a cavity resonator for tuning purposes. In the second approach, the shape of the cavity resonator is changed and as a consequence the resonance frequency is altered.



Mercier et al., “A micromachined tunable cavity resonator,”  
*Proceeding of the 33<sup>rd</sup> European Microwave Conference, June, 2003.*

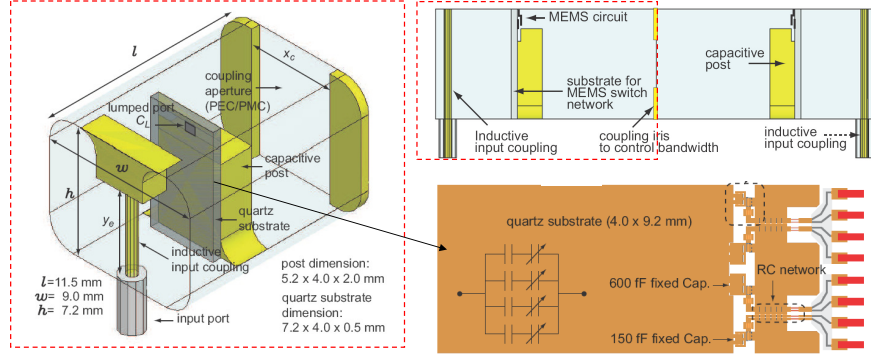
Fig. 1.9.: Silicon micromachined tunable (0.1 GHz) air cavity resonator at 26.8 GHz (Q-factor = 125) coupled to an external MEMS varactor [Mercier 03].

The first approach can be further categorized according to the location of the loading varactor i.e. inside or outside to the cavity. An early example of a tunable micromachined cavity resonator was presented by Mercier *et al.* at 26.8 GHz and corresponds to a micromachined cavity resonator externally loaded with a two-state MEMS varactor, as can be seen in Figure 1.9. A silicon micromachined air cavity resonator with a Q-factor of 125 and with a relatively low tuning range of 0.1 GHz at 26.8 GHz was demonstrated [Mercier 03].

The same principle is revisited by Li *et al.* at 60 GHz, but using this time a CPW line of different lengths to emulate a varactor. Actually, this is a static tuning demonstration as three different cavity resonators are compared. The resonators are characterized by a high Q-factor of 298 and a tuning range of 1.3 GHz (2.5 %) was demonstrated [Li 09].

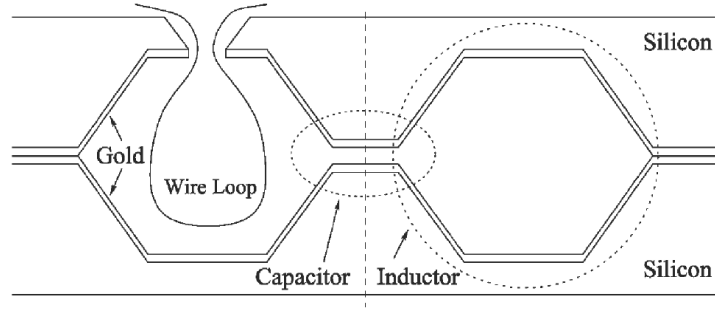
On the other hand, with the MEMS capacitor located inside the cavity resonator, Park *et al.* developed a tunable evanescent mode cavity filter using two cavity resonators incorporating MEMS varactors, see Figure 1.10. The MEMS varactors were metal-air-metal fixed scaling capacitor connected in series and switched by a 4-bit network. The ensemble covers a capacitance range of 160-630 fF, resulting in 4.3-5.5 GHz (24.5 %) frequency coverage of the filter, while maintaining high Q-factors for the resonators

of 273-511, [Park 09]. Similarly, Stefanini *et al.* used a two-state MEMS capacitor, located in the center of the cavity resonator, in a particular X-shaped post. This was resulting in a lower frequency tuning 4.96-5.54 GHz (11.6 %) as a single MEMS switch was used but the Q-factor was improved (552-862) [Stefanini 09].



Park *et al.*, “High-Q RF-MEMS 4-6 GHz tunable evanescent-mode cavity filter,” *IEEE Transactions on Microwave Theory and Techniques*, February, 2010.

Fig. 1.10.: High-Q (273-511) tunable (4.3-5.5 GHz, 24.5 %) evanescent mode cavity filter using capacitive RF-MEMS cantilever switches [Park 09].

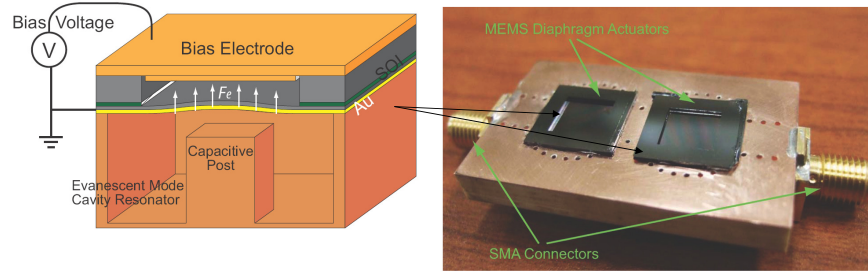


Hou *et al.*, “A High-Q widely tunable Gigahertz electromagnetic cavity resonator,” *Journal of Microelectromechanical Systems*, December, 2006.

Fig. 1.11.: Toroidal lumped cavity resonator, tunable (2.5 to 4.0 GHz, 46.15 %) with high Q-factor (120-240) [Hou 06].

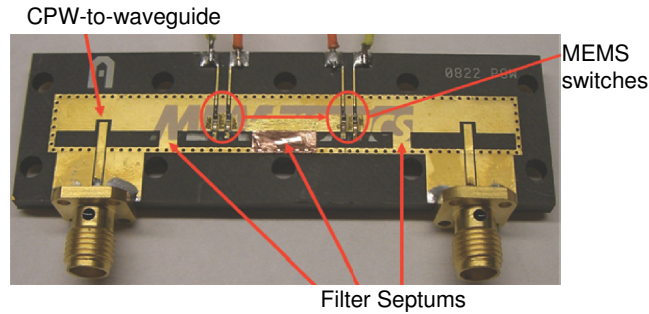
The second approach involves the deformation of some side walls of the cavity resonator, for tuning purposes. In this case, Hou *et al.* implemented a KOH-etched toroidal cavity lumped resonator, as the capacitive and inductive zones are well defined respectively in the center and on the side of the cavity, see Figure 1.11. The toroidal cavity resonator is squeezed in the center to make the capacitor variable and

therefore to change its resonance frequency. The actuation is realized using an external commercial piezoelectric actuator to widely tune a lumped cavity resonator, from 2.5 to 4.0 GHz (46.15 %) and a high Q-factor (120-240) is demonstrated [Hou 06].



Xiaoguang et al., “High-Q tunable microwave cavity resonators and filters using SOI-based RF MEMS tuners,” *Journal of Microelectromechanical Systems*, August, 2010.

Fig. 1.12.: High-Q (650-300) tunable (1.9-5 GHz, 88.57 %) microwave cavity resonators and filters using an electrostatically actuated SOI-based MEMS membrane [Xiaoguang 10].



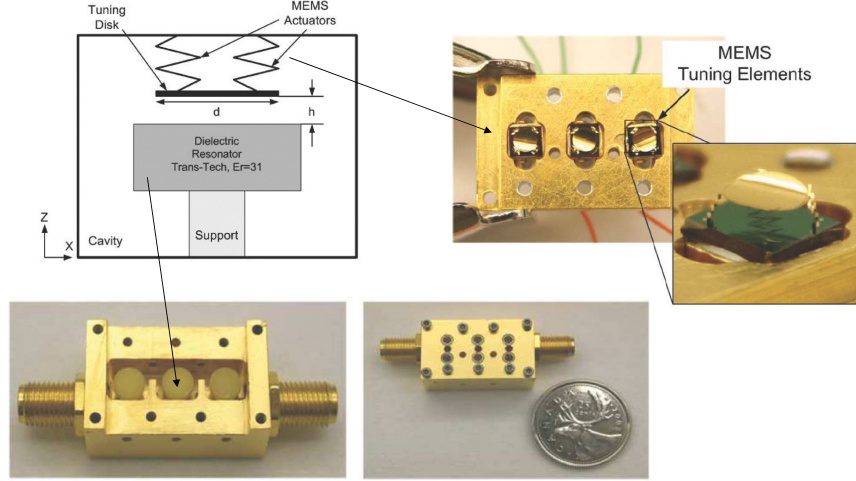
Scarbrough et al., “Miniature microwave RF MEMS tunable waveguide filter,” *IEEE European Microwave Integrated Circuits Conference*, September, 2009.

Fig. 1.13.: RF-MEMS tunable filter (10.12-10.33 GHz, 2.05 %) implemented in dielectric filled waveguide, Q-factors (169-319) [Scarbrough 09].

The best performing integrated cavity resonators, in terms of tunability while maintaining very high Q-factors (530-450), are realized by Xiaoguang *et al.* A high tuning range, from 3.4 to 6.2 GHz (58.33 %) was achieved by using a cavity resonator incorporating a perturbation post, above which an electrostatically actuated SOI MEMS membrane was actuated for tuning purposes, as can be seen in Figure 1.12, referring to [Xiaoguang 09]. Actually, the field distribution is characterized by a high electrical field in the center of the cavity resonator, where the post is located. As a consequence

the capacitance of the gap between the post and the membrane is high. Similar tunable cavity resonators were used in the design of evanescent tunable filters, demonstrating excellent tuning ranges (1.9-5 GHz, 88.57 %) and very high Q-factors (650-300) [Xiaoguang 10].

Alternatively, Scarbrough *et al.* use commuted studs patterned on standard PCB, introduced longitudinally in the center of a specially designed two parts metallic waveguide, filled with dielectric to change the filter's central frequency, as can be seen in Figure 1.13. RF-MEMS switches are used to couple/decouple these stubs. A large tuning range (10.12-10.33 GHz, 2.05 %) and high Q-factors (169-319) are demonstrated [Scarbrough 09].



Yan et al., "Tunable Dielectric Resonator bandpass filter with embedded MEMS tuning elements," *IEEE Transactions on Microwave Theory and Techniques*, January, 2007.

Fig. 1.14.: Tunable DR based passband filter (15.62-15.86 GHz, 1.52 %) with thermal MEMS actuators. Q-factors (1638-421) are demonstrated [Yan 07].

Finally, Yan and Mansour tuned a filter composed of three Dielectric Resonators, (DR) by MEMS thermally actuated metallic disks of  $2 \times 2 \text{ mm}^2$ , see Figure 1.14. A tuning range from 15.62 to 15.86 GHz (1.52 %) and very high Q-factors (1638-421) are demonstrated [Yan 07].

Table 1.1.: Chronological State of the Art of fixed in frequency and tunable integrated cavity resonators.

| ref.                 | Technology                                 | Center frequency (GHz) | $Q_0$    | size (mm <sup>3</sup> ) | Tuning range (GHz)    |
|----------------------|--------------------------------------------|------------------------|----------|-------------------------|-----------------------|
| [Papapolymerou 97]   | $\mu$ machined LR-Si                       | 10.285                 | 506      | 16×32×0.47              | -                     |
| [Harle 98, Harle 03] | $\mu$ machined LR-Si                       | 10.525                 | 506      | 16×32×0.47              | -                     |
| [Brown 99]           | $\mu$ machined LR-Si                       | 23.97                  | 1117     | 8.84×8.84×0.9           | -                     |
| [Kim 99]             | $\mu$ machined LR-Si                       | 30.14                  | 120      | 8.4×6.3×0.4             | -                     |
| [Stickel 01]         | $\mu$ machined LR-Si, multiple thicknesses | 30                     | 2155     | 10×10×2.5               | -                     |
| [El-Tager 03]        | LTCC process, air cavity                   | 19.8                   | 1206     | 8×9.9×0.7               | -                     |
| [Stickel 03]         | $\mu$ machined LR-Si, multiple thicknesses | 30                     | 4520     | 7.07×7.07×7.5           | -                     |
| [Harle 04a]          | $\mu$ machined LR-Si                       | 27                     | 1465     | 7.6×7.6×1               | -                     |
| [Ferrand 05]         | SIW in LTCC                                | 42.6                   | 350      | 1.45×1.45×0.4           | -                     |
| [Lee 05a]            | SIW in LTCC                                | 60                     | 367      | 1.95×1.3×0.3            | -                     |
| [Ma 07]              | LIGA process                               | 24                     | 2122     | 8.8×8.8×2               | -                     |
| [Ma 07]              | LIGA process                               | 37                     | 309      | 5.7×5.7×2               | -                     |
| [Posada 08]          | SIW in HR-Si                               | 29                     | 170      | 1.7×3.5×0.1             | -                     |
| [Posada 08]          | SIW in HR-Si                               | 60                     | 186      | 0.8×1.7×0.1             | -                     |
| [Ocket 09]           | $\mu$ machined LR-Si                       | 60                     | 744      | 3.8×3.8×0.46            | -                     |
| [Ocket 10]           | $\mu$ machined LR-Si                       | 74.16                  | 584      | 3.1×3.1×0.46            | -                     |
| [Mercier 03]         | $\mu$ machined LR-Si, ext. MEMS varicap    | 26.85                  | 125      | 7.6×7.6×0.525           | 26.8-26.9<br>0.37 %   |
| [White 05]           | metallic WG, piezo MEMS actuator           | 1.9                    | 380      | 14×14×3                 | 1.2-2.6<br>73.68 %    |
| [Hou 06]             | $\mu$ machined LR-Si                       | 3.25                   | 120-240  | 18×18×1                 | 2.5-4<br>46.15 %      |
| [Yan 07]             | DR in metallic WG, thermal actuator        | 15.74                  | 1638-421 | (5×5×3)                 | 15.62-15.86<br>1.52 % |
| [Li 09]              | $\mu$ machined LR-Si, fixed steps          | 60                     | 298      | 3.6×3.6×0.3             | 59.3-60.6<br>2.50 %   |
| [Xiaoguang 09]       | air cavity in Rogers, piezo MEMS actuator  | 4.8                    | 530-450  | 18×30×4                 | 3.4-6.2<br>58.33 %    |
| [Park 09, Park 10]   | metallic WG, MEMS switch                   | 4.9                    | 273-511  | 23×9×7.2                | 4.3-5.5<br>24.49 %    |
| [Scarborough 09]     | SIW in Duroid, MEMS switch                 | 10.225                 | 169-319  | 63.8×15×7.75            | 10.12-10.33<br>2.05 % |
| [Stefanini 09]       | metallic WG, MEMS switch                   | 5                      | 552-862  | 24×24×15                | 4.96-5.54<br>11.60 %  |
| [Xiaoguang 10]       | air cavity in Rogers, piezo MEMS actuator  | 3.5                    | 650-300  | 30×18×4.5               | 1.9-5<br>88.57 %      |

## 1.2 RESEARCH OBJECTIVES

The research on tunable integrated cavity resonators, see Table 1.1 is relatively new, with a first publication in 2003 by [Mercier 03]. Diverse technologies are supporting these realizations, very often the demonstrators are using standard metallic waveguides in association with MEMS [White 05, Park 09, Park 10, Stefanini 09, Yan 07]. The integration exists for the frequencies 1.9 to 15.74 GHz, as reported in Table 1.1. For higher frequencies, the manual handling of small components is difficult or impossible and always prohibitive. Alternatively, air cavity resonators shaped by vias through substrate, incorporating a central post and a MEMS membrane demonstrated the best performance to date, in terms of tunability and Q-factor [Xiaoguang 10, Xiaoguang 09]. Similarly, the central post was commuted in dielectric filled cavity resonators [Scarborough 09]. Micromachined air cavity resonators in silicon were first realized by [Mercier 03] and [Hou 06]. This technology is recommended at higher frequencies. Back in 2006, at the beginning of this PhD, an integrated tunability system for micromachined cavity resonators at 60 GHz was missing. A demonstration at this frequency was published in 2009, but consists in a static comparison of three different cavity resonators [Li 09]. One of the main results of this thesis is the realization of a functional voltage controlled tunable integrated resonator, at 60 GHz [Dancila 10]. The other objectives of this thesis are summarized as follows:

Investigate the characterization techniques and modelization of fixed and tunable in frequency integrated cavity resonators.

Integrate a tuning system and realize a high Q-factor silicon micromachined air cavity resonator, at 60 GHz. Important issues appearing during the integration process, like the packaging of the components while preserving MEMS integrity are hard barriers of this manufacture. The fabrication of the devices was not realized by the author, nevertheless a close focus on the technological development was necessary to obtain the successful results of this PhD.

Investigate the perturbation of the cavity resonator's non-homogeneous field distribution by non-homogeneous volumes of perturbation, for tuning and miniaturization purposes. Actually, the tuning approach commonly followed is reduced to the variation of the capacitor of the cavity resonator's lumped elements, LC. Opportunities offered by a more in depth analysis were commonly neglected because of the higher complexity required. The impact of such perturbations, on the resonance frequency and on the Q-factor was never fully investigated.

Finally, as an extension of the research on integrated cavity resonators, improve the State of the Art by realizing passives and actives applications. The passive application realized is a seven-pole filter and the active is an ultra-low phase-noise oscillator, both improving the State of the Art by using high-Q substrate integrated cavity resonators at 60 GHz.

### 1.3 THESIS OUTLINE

The research objectives, are met in this thesis and the results are presented in six chapters, as follows:

**Chapter 2.** As a preliminary discussion, I revisit the foundations of cavity resonators.

Theoretical aspects are presented and practical characterization techniques are disclosed. An important aspect in the study of cavity resonators is constituted by the coupling mechanism. Therefore, I study different kinds of couplings and investigate the theory underlying their development towards a confrontation with practical realizations. The concept of transfer of energy, using the definition of the effective volume is used to evaluate the coupling coefficient of a microstrip line feed. The impedance resulting the field distribution is further investigated, in a similar way to the patch antenna analysis. I use this concept in the design of a cavity resonator with a coplanar feed. As commonly a mix of both impedance variation and energy transfer occurs during the feeding of any cavity resonator, feeds mixing those two aspects are also investigated.

In the second part, I investigate a direct application of these concepts for the integration of a cavity resonator in the bulk silicon of IMEC's MCM-D, at 60 and 75 GHz. A more detailed analysis follows, emphasizing the extraction of the unloaded resonance frequency of the cavity resonator which is otherwise changed by any coupling system. The unloaded frequency is used in the establishment of very accurate lumped equivalent circuits.

Finally, I present a new miniaturization concept for cavity resonators with the aim to open a new perspective in this domain. The idea is to create a quarter wavelength long cavity resonator by using a combination of Perfect Electric Conductor (PEC) and Perfect Magnetic Conductor (PMC) boundary conditions. As a consequence, the side dimensions are reduced by a factor two and the volume is reduced by a factor four. The practical implementation of the PMC is realized using High Impedance Surfaces (HIS). However, the integration of this concept was not realized within this thesis. Nevertheless, alternative ways to miniaturize cavity resonators are derived from the tuning and trimming investigation presented in Chapter 3 and for which the integration was realized.

**Chapter 3.** This chapter focuses on the study of tuning and trimming concepts for air cavity resonators. I evaluate analytically the resonance frequency and the Q-factor of a cavity resonator incorporating internal volumes of perturbation. The perturbation is either located in the center of the cavity resonator, called Central Post Perturbation (CPP) or dually, a hole shaped in an annular extruded volume, called Side Annular Perturbation (SAP). These particular shapes derive from the duality of the field distribution present in the cavity resonator, where the electric field is maximal in the center and minimal on the edges of the cavity, while it is the opposite for the magnetic field. I confront the theory to practical realizations, and therefore use 72 prototypes of internally perturbed resonators, addressing both kinds of perturbations. Each prototype is characterized by different lateral dimensions and different heights, sweeping from the idle empty case, to the fully filled one. The theory is in very good agreement with the measurements.

In the second part, I design an integrated tunable air cavity resonator at 60 GHz, based on MEMS cantilevers for tuning purposes. This realization uses a new concept, developed expressly for integrated and tunable air cavity resonators. This is in fact a Faraday cage composed of up-curved cantilevers located inside a silicon micromachined cavity resonator. The Faraday cage acts similarly to the volume of perturbation previously studied and perturbs the electromagnetic field distribution inside the cavity resonator, tuning the resonance frequency of the resonator. The fabrication process is described and the successive designs are reviewed.

As several issues limit the Q-factors of the realized resonators, but not the tunability, I conduct the investigation further to identify the sources of the mentioned deterioration. As a result of this investigation, a high Q-factor air cavity resonators was successfully realized at 74.16 GHz.

The fourth and fifth chapters are a direct application of the previous chapters. They share with the rest of the thesis the design of a cavity resonator for a given frequency, in a given technology and the development of an adequate feeding.

**Chapter 4.** In this chapter I present a large bandwidth filter of 7 GHz around a central frequency of 60 GHz. Following the specifications of the filter in terms of passband' bandwidth, rejection outside the passband and insertion loss, a seven-pole is required. The filter is composed of seven cavity resonators, each being coupled by inductive irises. I designed the filter using tabulated element values for Chebyshev multi-pole filters, providing the values of the coupling coefficients between the filter's resonators as well as the values of the external couplings. The filter is fully simulated, optimized and compared to measurements.

**Chapter 5.** In this chapter I investigate the realization of an ultra-low phase-noise oscillator at 60 GHz. This oscillator is based on an integrated high Q-factor cavity resonator. Actually, using a high Q-factor integrated cavity resonator instead of integrated passive components, the latest are limited in performance at mm-wave, improves oscillator's performance. Furthermore, the demand for such oscillators is increasing as for example automotive RADAR for collision prevention systems are sought to widely be introduced by cars manufacturers. Similar devices, already on the market are available at prohibitive costs. The successful realization of an ultra-low phase noise oscillator imposes a very good comprehension of the operational principles and design techniques. This is to avoid the generation of spurious oscillations at unwanted frequencies and to reduce phase noise.

As a result of two successive iterations, this Chapter is divided in two parts: the first part assesses the design, characterization and realization of such an oscillator in MCM-D; the second part, is about a similar realization in LTCC. Even though, the first realization is presenting fair results, better performance was expected. Actually, the improvement of the LTCC iteration over the MCM-D one is not exactly due to the better efficiency of the LTCC technology over the MCM-D's, but to the operation of the cavity resonator at the resonance frequency, which



was missing in the MCM-D's implementation, suffering from some phase shift discrepancies.

Finally, the results, presented at the end of this thesis demonstrate the best performance as to our knowledge and to date, improving the State of the Art of integrated oscillators in the 40 to 80 GHz frequency band. A measured phase noise of -116 dBc/Hz @ 1 MHz offset of the carrier and an output power,  $P_{OUT} = 10.03$  dBm are reported at the oscillation frequency  $f_{osc} = 59.98$  GHz.

**Chapter 6.** This chapter is concluding and introduces further research directions for future work based on the premises of this thesis.



## CHAPTER 2

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# CAVITY RESONATORS: GENERAL THEORY, MODELLING AND CHARACTERIZATION

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### 2.1 INTRODUCTION

Metallic rectangular and cylindrical cavity resonators have received a large amount of interest during the history of radio telecommunications, as they offer very high performance in terms of frequency selectivity [Collin 66, Harrington 01, Pozar 05]. These resonators are largely covered by theoretical developments and lectures as for example by Feynman in his famous *"Lectures on Physics"* [Feynman 64] and more practically by Pozar in *"Microwave Engineering"* [Pozar 05].

In the industry, metallic cavity resonators are commonly used to realize RF filters [Boria 07] and oscillators [White 05], especially designed for satellite and military telecommunications. Even though the metallic cavity resonators can be quite bulky, high in weight, large in size, costly to manufacture and difficult to integrate with electronic circuits, the interest in using such devices resides in the high Q-factors they provide. In general for a filter, the higher the resonator's Q-factor, the lower the level of insertion loss for a given selectivity. Likewise, for an oscillator, the higher the resonator's Q-factor, the lower the phase noise.

This chapter covers first the theoretical foundations of resonators, addressing general considerations about lumped RLC resonators and focusing on the Q-factors extraction. Next, the feed design is investigated and a large variety of prototypes are characterized at 4 GHz. The characterization methods are assessed and an essential understanding of the coupling mechanism is provided. For this purpose, analytical developments are compared to full wave simulations and measurements for both rectangular and cylin-

drical cavity resonators. Based on this knowledge, the integration of cylindrical cavity resonators at 60 and 75 GHz is successfully realized in IMEC's MCM-D integration technology. State of the art performance is achieved by resonators with characteristic unloaded Q-factors:  $Q_0 = 117$  and  $Q_0 = 91$  respectively at 60 GHz and 75 GHz. Additionally, the characterization methodology is improved by developing the extraction of equivalent circuits. This model is based on the unloaded resonance frequency, extracted from a family of cavity resonators wherein each cavity is characterized by a different coupling coefficient. Finally, a miniaturization concept for air cavity resonators is demonstrated, with the aim to open a new perspective in this domain. The concept consists in shaping a quarter wavelength long cavity resonator by using a combination of Perfect Electric Conductor (PEC) and Perfect Magnetic Conductor (PMC) boundary conditions. As a consequence, the side dimensions are reduced by a factor two and the volume is reduced by a factor four. The practical implementation of the PMC is realized using High Impedance Surfaces (HIS).

## 2.2 LUMPED RESONATOR

The resonators are circuits that store energy and cyclically convert the nature of the stored energy, i.e. from electric to magnetic energy. Unfortunately, the storage is not perfect but generally losses are low. Resonators can be either lumped or distributed. The main distinction between these two is that the lumped element circuits have a finite number of poles and zeros, following the number of lumped component resonators composing the network while the microwave circuits have an infinite number of them. This is summarized from Kajfez's excellent book "*Q-factor*", inspiring this section [Kajfez 94].

In the following pages, we focus on the analysis of RLC lumped resonators with capacitor ( $C$ ), inductor ( $L$ ) and where losses, represented by the resistor ( $R$ ) are taken into account. In its simple form, the resonator can take either a serial or a parallel form. We start with those simple lumped-elements circuits, as the operation of microwave resonators is very similar.

### 2.2.1 Series RLC resonator

A typical series resonant circuit is represented in Figure 2.1. To analyze the circuit we consider the common current,  $I$  which allows writing the input impedance:

$$Z_s = R_s + j \left[ \omega L - \frac{1}{\omega C} \right] \quad (2.1)$$

At the resonance frequency, the imaginary part of the impedance is equal to zero and  $\omega_0 = 2\pi f_0$  is given by:

$$\omega_0 = \frac{1}{\sqrt{LC}} \quad (2.2)$$

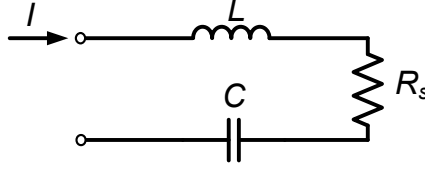


Fig. 2.1.: Series RLC resonant circuit.

The definition (2.2) allows us to rewrite (2.1) as:

$$Z_s = R_s \left[ 1 + j \frac{\omega_0 L}{R_s} \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) \right] \quad (2.3)$$

Considering the behavior of the circuit at resonance, assuming  $\omega = \omega_0 + \Delta\omega$  with  $\Delta\omega \ll \omega_0$ , (2.3) is approximated by:

$$Z_s \approx R_s \left[ 1 + j 2Q_{0s} \frac{\omega - \omega_0}{\omega_0} \right] \quad (2.4)$$

where  $Q_{0s}$  is the unloaded Q-factor of the serial RLC circuit:

$$Q_{0s} = \frac{\omega_0 L}{R_s}. \quad (2.5)$$

Considering further the normalized frequency difference,  $\xi$ :

$$\xi = 2Q_{0s} \frac{\omega - \omega_0}{\omega_0} = 2Q_{0s} \delta \quad (2.6)$$

where  $\delta = \frac{\omega - \omega_0}{\omega_0}$ , the impedance (2.4) can be rewritten as:

$$\frac{Z_s}{R_s} \approx 1 + j\xi \quad (2.7)$$

The normalized admittance,  $Y_s R_s$  is the inverse of (2.7):

$$Y_s R_s \approx \frac{1}{1 + j\xi} \quad (2.8)$$

### 2.2.2 Parallel RLC resonator

A typical parallel resonant circuit is represented in Figure 2.2. To analyze the circuit we consider the common voltage,  $V$  which allows writing the admittance:

$$Y_p = \frac{1}{R_p} + j \left[ \omega C - \frac{1}{\omega L} \right] \quad (2.9)$$

The definition (2.2) allows us to rewrite also (2.9) as:

$$Y_p = \frac{1}{R_p} \left[ 1 + j\omega_0 C R_p \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) \right] \quad (2.10)$$

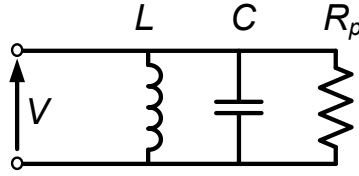


Fig. 2.2.: Parallel RLC resonant circuit.

In the vicinity of the resonance, (2.10) is approximated by:

$$Y_p = \frac{1}{R_p} \left[ 1 + j2Q_{0_p} \frac{\omega - \omega_0}{\omega_0} \right] \quad (2.11)$$

where  $Q_{0_p}$  is the unloaded Q-factor of the parallel RLC circuit:

$$Q_{0_p} = \omega_0 C R_p. \quad (2.12)$$

Considering further the normalized frequency difference  $\xi$ , (2.6), the admittance (2.11) can be normalized as:

$$Y_p R_p \approx 1 + j\xi \quad (2.13)$$

The normalized impedance,  $\frac{Z_p}{R_p}$  is the inverse of (2.13):

$$\frac{Z_p}{R_p} \approx \frac{1}{1 + j\xi} \quad (2.14)$$

### 2.2.3 Duality of the parallel and series RLC resonator

The admittance of the series RLC resonator is similar to the impedance of the parallel RLC resonator, as can be seen in Figure 2.3 and the impedance of the series RLC resonator is similar to the admittance of the parallel RLC resonator, as can be seen in Figure 2.4. This is due to the fact that both circuits are dual under the voltage, current permutation.

Around the resonance, the real and imaginary part of the normalized admittance of the series RLC resonator (2.1) are shown in Figure 2.3. They are similar to the real and imaginary part of the normalized impedance of the parallel RLC resonator (2.9). The real part of the admittance of the series RLC (impedance of the parallel RLC) has a maximum at the resonance frequency ( $\xi = 0$ ). The imaginary part shows a

maximum ( $\xi = -1$ ) and a minimum ( $\xi = 1$ ), in between the slope of the susceptance is negative contradicting Foster's theorem [Foster 24], as a consequence of additional losses.

The real and imaginary part of the normalized admittance of the series lossless LC resonator of impedance:  $Z_{0s} = j [\omega L - \frac{1}{\omega C}]$  are also presented. Similarly, they are compared to the real and imaginary part of the normalized impedance of the parallel lossless LC resonator of admittance:  $Y_{0p} = j [\omega C - \frac{1}{\omega L}]$ . These are compared to the equations (2.8) and (2.14) of the narrow band approximation of the admittance of the series RLC and impedance of the parallel RLC, respectively.

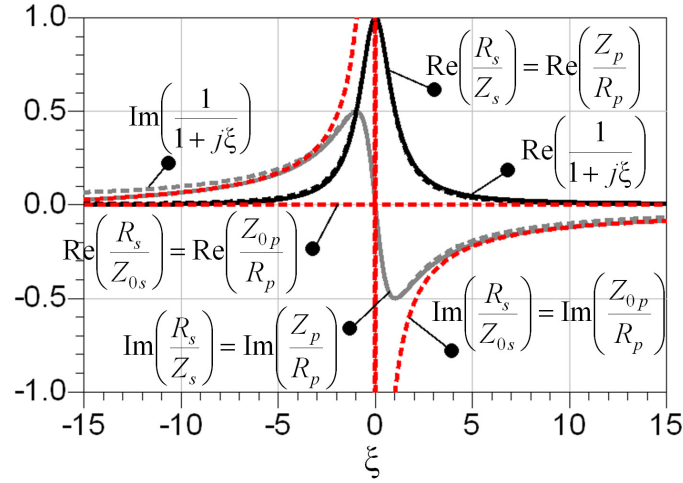


Fig. 2.3.: Real and imaginary parts of the normalized in magnitude admittance of a series resonant circuit, magnitude of the impedance of a parallel resonant circuit and narrow band approximations (2.8) and (2.14), as a function of  $\xi$ .

Around the resonance, the real and imaginary part of the normalized impedance (2.1) of the series RLC resonator are shown in Figure 2.4. They are similar to the real and imaginary part of the normalized admittance of the parallel RLC resonator (2.9). The imaginary part of the admittance of the parallel RLC (impedance of the series RLC), a nonlinear function of frequency, is approximated by a straight line passing through zero at the resonance frequency ( $\xi = 0$ ). The approximation is only valid in a narrow band around the resonance and therefore the description of a microwave circuit in terms of Q-factor is a narrow-band approximation.

The real and imaginary part of the normalized impedance of the series lossless LC resonator of impedance:  $Z_{0s} = j [\omega L - \frac{1}{\omega C}]$  are also presented. They are also compared to the real and imaginary part of the normalized admittance of the parallel lossless LC resonator of admittance:  $Y_{0p} = j [\omega C - \frac{1}{\omega L}]$ . These are compared to the equations (2.7) and (2.13) of the narrow band approximation of the impedance of the series RLC and admittance of the parallel RLC, respectively.

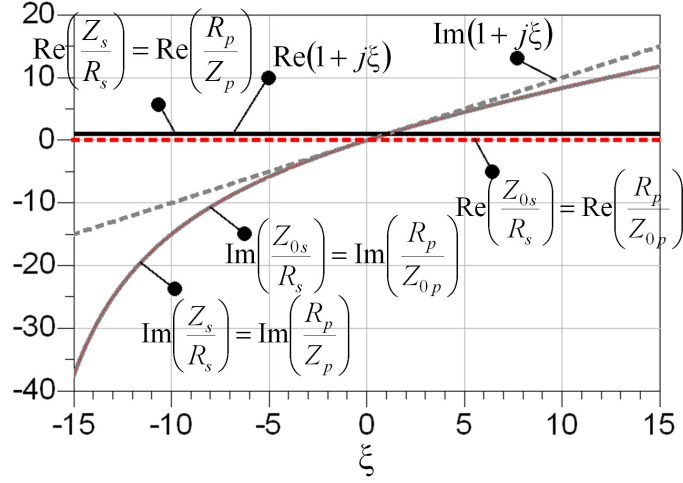


Fig. 2.4.: Real and imaginary parts of the normalized in magnitude impedance of a series resonant circuit, magnitude of the admittance of a parallel resonant circuit and narrow band approximations (2.7) and (2.13), as a function of  $\xi$ .

## 2.3 Q-FACTOR

In this Section, the definitions of the unloaded or unloaded Q-factor and coupling coefficient are introduced. The fundamentals of the analytical extraction of the Q-factors are developed for one port resonators. A convenient method for extraction is presented as implemented in the commercially available software QZERO.

### 2.3.1 Unloaded Q-factor definition

For lumped-elements resonators, as well as for microwave resonators, the unloaded Q-factor,  $Q_0$  is the energy ratio between the maximum energy stored in the resonator, the peak value of the stored energy,  $W_{max}$  to the dissipated energy in the resonator during one period,  $W_d$ . The ratio is evaluated at the resonance frequency,  $\omega_0$  [Kajfez 94]:

$$Q_0 = 2\pi \left[ \frac{W_{max}}{W_d} \right]_{\omega=\omega_0} \quad (2.15)$$

Considering the period of oscillation is  $T = \frac{2\pi}{\omega}$  the energy dissipated in one period is:  $W_d = P_d T$ , where  $P_d$  is the average power dissipated in the resonator. Thus (2.15) can be rewritten as:

$$Q_0 = \left[ \omega \frac{W_{max}}{P_d} \right]_{\omega=\omega_0} \quad (2.16)$$



Considering the parallel RLC resonator in Figure 2.2 with common voltage:  $v(t) = |V| \cos(\omega t)$ , the corresponding stored electric energy in the capacitor is:

$$W_e(t) = \frac{1}{2} C v(t)^2 = \frac{1}{2} C |V|^2 \cos^2(\omega t) \quad (2.17)$$

The current through the inductor is then  $i_L(t) = \frac{1}{L} \int v(t) dt = \frac{1}{\omega L} |V| \sin(\omega t)$  and the magnetic energy stored in the inductor is derived as follows:

$$W_m(t) = \frac{1}{2} L i(t)^2 = \frac{1}{2} \frac{1}{\omega^2 L} |V|^2 \sin^2(\omega t) \quad (2.18)$$

Both stored electric and magnetic energies evolve in time at double the resonance frequency as can be deduced from 2.17 and 2.18. This means that one period of  $v(t)$  or  $i(t)$  corresponds to two periods of  $W_e(t)$  or  $W_m(t)$ . Furthermore, the maximum of the electric energy stored  $W_e(t) = W_{e,max}$  happens at the minimum of the magnetic energy stored  $W_m(t) = 0$  and *vice versa*. Considering that at resonance, i.e.  $\omega = \omega_0$ , the average values  $W_e$  and  $W_m$  are equal to one-half of their peak values, the stored energies to be used in (2.16) are, indifferently:

$$W_{max} = 2W_e = 2W_m = W_e + W_m \quad (2.19)$$

therefore (2.16) can be rewritten in terms of the average stored energies as:

$$Q_0 = \left[ \omega \frac{W_e + W_m}{P_d} \right]_{\omega=\omega_0} \quad (2.20)$$

Following a development that we do not detail here but that could be found in [Kajfez 94], the Q-factor can also be expressed in terms of the slope of the reactance of a series resonator, as a function of frequency:

$$Q_0 = \left[ \frac{\omega}{2R_s} \frac{dX}{d\omega} \right]_{\omega=\omega_{0,s}} \quad (2.21)$$

where the power is only dissipated in the equivalent resistance  $R_s$  and for a parallel resonator, the Q-factor can also be expressed in terms of the slope of the susceptance, as a function of frequency:

$$Q_0 = \left[ \frac{\omega}{2G_p} \frac{dB}{d\omega} \right]_{\omega=\omega_{0,p}} \quad (2.22)$$

where the power is only dissipated in the equivalent conductance  $G_p = 1/R_p$ .

The power dissipated in conductive losses due to the conductivity of the cavity's walls is called:  $P_{cond}$ . The power dissipated in the lossy dielectric filling the cavity resonator is called:  $P_{diel}$ . Additionally, any opening in the resonator enclosure results in an additional radiative losses, called  $P_{rad}$ . Therefore the total power dissipated is a summation of partial sources of loss  $P_d = P_{cond} + P_{diel} + P_{rad}$  and therefore the

unloaded Q-factor,  $Q_0$  is a summation of partial Q-factors accounting for the sources of loss, as follows:

$$\frac{1}{Q_0} = \frac{1}{Q_{cond}} + \frac{1}{Q_{diel}} + \frac{1}{Q_{rad}} \quad (2.23)$$

where  $Q_{cond}$  is the unloaded quality factor when only conductive losses are considered,  $Q_{diel}$ , the unloaded quality factor when only dielectric losses are considered and  $Q_{rad}$ , the unloaded quality factor when only radiation losses are considered.

### 2.3.2 Loaded Q-factor definition

When the unloaded resonator is connected, a coupling coefficient,  $\kappa$  is defined as the ratio of power dissipated in the external circuit,  $P_e$  to the power dissipated in the resonator,  $P_0$ :

$$\kappa = \frac{P_e}{P_0} = \frac{Q_0}{Q_e} \quad (2.24)$$

The expression (2.24) is best illustrated in the parallel resonator, considering common voltage. It follows that  $P_e = V^2/R_c$  and  $P_0 = V^2/R_p$  and therefore  $\kappa = R_p/R_c$ , the resistance normalized to the load charge resistance  $R_c$ , following Figure 2.5. When  $\kappa < 1$  the coupling is called undercoupled and the losses in the external circuit are lower than in the cavity. When  $\kappa > 1$  the coupling is called overcoupled and the power dissipated in the external circuit is larger in comparison with the power dissipated in the unloaded cavity. When  $\kappa = 1$  the coupling is critical and external and cavity losses are equal. Despite very high values the unloaded Q-factor could reach, when the resonator is connected, the resulting quality factor is the loaded Q-factor, which is lowered by the losses in the coupling mechanism. Actually, the loaded Q-factor,  $Q_l$  is lowered in comparison with  $Q_0$  due to the additional  $R_c$  in parallel with  $R_p$ , as can be seen in Figure 2.5.

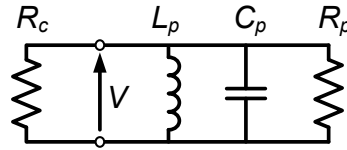


Fig. 2.5.: Loaded lumped parallel RLC resonant circuit.

The relation between the external Q-factor,  $Q_e$ , the loaded Q-factor,  $Q_l$  and unloaded Q-factor,  $Q_0$  is given as [Pozar 05]:

$$\frac{1}{Q_l} = \frac{1}{Q_0} + \frac{1}{Q_e} \quad (2.25)$$

Finally, using (2.24) and (2.25) the loaded Q-factor is rewritten as a function of the coupling coefficient and of the unloaded Q-factor:

$$Q_l = \frac{1}{1 + \kappa} Q_0 \quad (2.26)$$

The unloaded Q-factor is the asymptotic value the loaded Q-factor would ever reach. If reached the resonator is not coupled. Eventually, the coupling can take different values as required by the circuitry using the resonator. A remarkable value appears when the resonator shows a matched impedance to the external world, in the critically coupled case, when the loaded Q-factor is half the unloaded Q-factor.

### 2.3.3 Analytical extraction of the Q-factors

The extraction of  $Q_0$ ,  $Q_e$  and  $Q_l$  is obtained from the impedance or admittance response of the resonator, by identification of three specific points, called the three points method. The ratio is taken between the resonance frequency and  $\Delta f = f_2 - f_1$ , the frequency bandwidth between two other points. These points are selected on the intersection with the resonance locus of six pre-defined loci, characteristic for the three Q-factors. Therefore, the unloaded Q-factor is evaluated as follows. Considering the normalized impedance as a function of the Q-factor, for a parallel RLC resonator (2.8) and the coupling coefficient at resonance:

$$\beta = \kappa(f_0) = \frac{R_p}{R_c} \quad (2.27)$$

allows re-normalized (2.14) to the characteristic impedance,  $R_c$ :

$$z = \frac{Z}{R_c} \approx \frac{\beta}{1 + j\xi} \quad (2.28)$$

When  $\xi = \pm 1$ , at frequencies  $f_{\pm Q_0}$ , the normalized impedance becomes:

$$z(f_{\pm Q_0}) \approx \frac{\beta}{1 \pm j} = \frac{\beta}{2}(1 \mp j) \quad (2.29)$$

and following (2.6), at frequencies  $f_{\pm Q_0}$ , the unloaded Q-factor is given by:

$$Q_0 = \frac{f_0}{2|f_{\pm Q_0} - f_0|} \quad (2.30)$$

Therefore in order to calculate  $Q_0$ , the frequencies  $f_{+Q_0}$  and  $f_{-Q_0}$  are extracted from the impedance satisfying the equations:

$$\text{Re}\{z(f_{\pm Q_0})\} = \mp \text{Im}\{z(f_{\pm Q_0})\} \quad (2.31)$$

The other Q-factors are derived from  $Q_0$  using (2.26) and (2.24) as:

$$Q_e = \frac{Q_0}{\beta} \quad (2.32)$$

$$Q_l = \frac{Q_0}{1 + \beta} \quad (2.33)$$

Similarly, the external Q-factor is extracted first and the other Q-factors are derived. For this purpose, (2.28) is rewritten in its admittance form, as follows:

$$y = Y R_C \approx \frac{1}{\beta}(1 + j\xi) \quad (2.34)$$

Considering (2.6) and (2.24), resulting in  $\xi = 2\beta Q_e \delta$ , the external Q-factor is also given by:

$$Q_e = \frac{f_0}{2|f_{\pm Q_e} - f_0|} \quad (2.35)$$

When  $\xi = \pm \beta$ , the normalized admittance becomes:

$$y(f_{\pm Q_e}) \approx \left(\frac{1}{\beta} \pm j\right) \quad (2.36)$$

Therefore in order to extract  $Q_e$ , the frequencies  $f_{+Q_e}$  and  $f_{-Q_e}$  are identified in the admittance satisfying the equations:

$$\text{Im}\{y(f_{\pm Q_e})\} = \pm 1 \quad (2.37)$$

The other Q-factors are derived from  $Q_e$  using (2.26) and (2.24) as:

$$Q_0 = \beta Q_e \quad (2.38)$$

$$Q_l = \frac{\beta}{1 + \beta} Q_e \quad (2.39)$$

Similarly, the loaded Q-factor can be extracted first and the other Q-factors derived. Considering (2.6) and (2.26), resulting in  $\xi = 2(1 + \beta)Q_l \delta$ , the loaded Q-factor is rewritten as:

$$Q_l = \frac{f_0}{2|f_{\pm Q_l} - f_0|} \quad (2.40)$$

when  $\xi = \pm (1 + \beta)$  the normalized admittance (2.34) is:

$$y(f_{\pm Q_l}) \approx \frac{1}{\beta} + j\left(\frac{1}{\beta} \pm 1\right) \quad (2.41)$$

Therefore in order to calculate  $Q_l$ , the frequencies  $f_{+Q_l}$  and  $f_{-Q_l}$  are extracted from the admittance satisfying the equations:

$$\text{Re}\{y(f_{\pm Q_l}) \pm 1\} = \pm \text{Im}\{y(f_{\pm Q_l})\} \quad (2.42)$$

The other Q-factors are derived from  $Q_l$  using (2.26) and (2.24) as:

$$Q_0 = (1 + \beta)Q_l \quad (2.43)$$

$$Q_e = \frac{1 + \beta}{\beta}Q_l \quad (2.44)$$

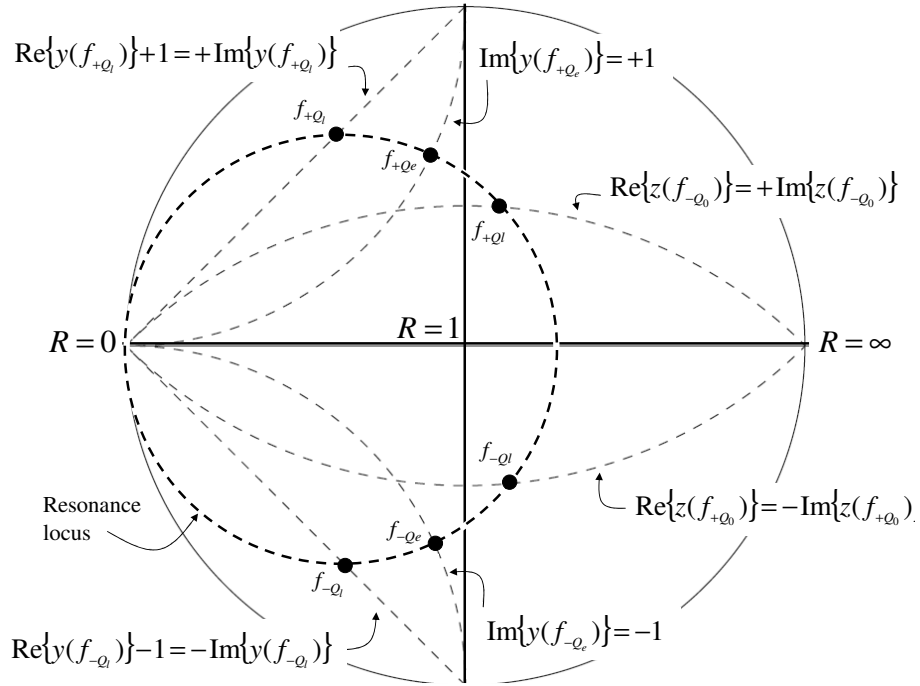


Fig. 2.6.: Identification of the characteristic points loci on the Smith chart of the three points method.

In the three points method, the Q-factors are obtained indifferently from the analytical method of extraction used, i.e. starting first with  $Q_0$ ,  $Q_e$  or  $Q_l$ . As shown in Figure 2.6, the two loci required to extract  $Q_0$  are two circles on the Smith chart, with the center at  $\pm 90^\circ$  and passing through 0 and  $180^\circ$  on the unitary circle; the two loci required to extract  $Q_e$  are two circles on the Smith chart, with the susceptance equal to one and the two loci required to extract  $Q_l$ , are two circles on the Smith chart, with the susceptance equal to the conductance plus one.

However, a major drawback of this method is that it requires to rearrange the measured data relative to the reference RLC resonator and therefore to remove the reactance of the coupling circuit. This is complicated as the reactance is not known in

advance. Furthermore, this method cannot handle effectively lossy measurement, i.e. resonance loci detached from the unitary edge of the Smith chart. Finally, to rectify these two points and to improve accuracy, additional data can be taken into account and interpolation is realized between the measurement points.

### 2.3.4 Q-factors extraction with QZERO

In this thesis, practically, the extraction of Q-factors from measurements is realized using the QZERO program, which is distributed with Kajfez's book "*Q-factor*" [Kajfez 94]. This choice is supported by the fact that QZERO alleviate all three drawbacks of the analytical method exposed in the previous Section 2.3.3, as it does not require to remove the reactance of the coupling circuit, can handle lossy circuits and improves the accuracy of the calculations as it realizes a linear fractional curve-fitting, based on which the Q-factors are extracted. It allows additionally, the calculation of random, systematic and accidental errors.

To realize the extraction the QZERO program is implementing an overdetermined measurement procedure, in which at least 20 points are used with data processing to extract the unloaded Q-factor, the coupling coefficient and the unloaded frequency of resonance. The functioning of the program is briefly described here below and more details can be found in [Kajfez 94].

In the lossless case, the detuned reflection coefficient is unitary in magnitude ( $|\Gamma_d| = 1$ ), as presented in Figure 2.7 and is given by:

$$\Gamma_d = e^{-j2\delta} \quad (2.45)$$

The input reflection coefficient  $\Gamma_i$  can be rewritten as:

$$\Gamma_i = \Gamma_d + \frac{D_1 e^{-j2\delta}}{1 + jQ_l 2 \frac{f-f_r}{f_0}} \quad (2.46)$$

where  $D_1$  is the diameter of the resonance locus and  $\kappa$  is the coupling coefficient:

$$D_1 = \frac{2\kappa}{1 + \kappa} \quad (2.47)$$

$\Gamma_i$  is reduced to the standard linear fractional transformation, suitable for data fitting:

$$\Gamma_i = \frac{a_1 t + a_2}{a_3 t + 1} \quad (2.48)$$

where  $a_1$ ,  $a_2$  and  $a_3$  are the transformation coefficients and  $t$  is the fractional linear transformation of the normalized frequency:

$$t = 2 \frac{f - f_r}{f_r} \quad (2.49)$$

The input coefficient at  $t = 0$ , when  $(f = f_r)$  is denoted  $\Gamma_r$ .  $\Gamma_r$  is the point on the resonance locus which is the closest to the origin, as can be seen in Figure 2.7. When  $t$  tends to infinity  $\Gamma_d = \frac{a_1}{a_3}$ . The mapping is from  $t$  to  $\Gamma_i$ , with the following constants:

$$a_1 = jQ_l\Gamma_d \quad (2.50)$$

$$a_2 = \Gamma_d + D_1e^{-j2\delta} \quad (2.51)$$

$$a_3 = jQ_l \quad (2.52)$$

Therefore, the coupling coefficient and loaded quality factor are derived as follows:

$$Q_l = \text{Im}(a_3) \quad (2.53)$$

$$\kappa = \frac{1}{\frac{2}{D_1} - 1} \quad (2.54)$$

with

$$D_1 = \left| a_2 - \frac{a_1}{a_3} \right| \quad (2.55)$$

Finally, the unloaded Q-factor is computed as (2.26):

$$Q_0 = Q_l(1 + \kappa) \quad (2.56)$$

Three reflection coefficient points at different measured frequencies are enough to extract the three transformation constants. Instead  $N$  measurement points are considered with (2.48) and the following overdetermined systems is established:

$$a_1t_n + a_2 - a_3t_n\Gamma_{i,n} = \Gamma_{i,n} \quad n = 1, 2, \dots, N \quad (2.57)$$

The overdetermined system of equations is then solved in a least-squares sense, weighted by weights inversely proportional to the variances of each measurement [Kajfez 94]. These weights suppress the influence of less reliable measured data points. The curve fitting allows to extract  $a_1$ ,  $a_2$  and  $a_3$  from  $N$  input reflection coefficients measured at  $N$  different frequencies.

A difficulty still remains in the identification of the loaded ( $f_r$ ) and unloaded resonance frequency ( $f_0$ ). Therefore, the procedure to determine  $f_r$  is iterative. In the first iteration, the loaded ( $f_r$ ) and unloaded resonance frequency ( $f_0$ ), used in fractional linear transformation of the normalized frequency (2.49), is taken as the frequency of the resonance locus of which point is the closest to the origin of the Smith chart, i.e. it has the smallest reflection coefficient,  $\Gamma_r$ . The center of the resonance locus,  $\Gamma_c$ , its diameter  $D_1$  and the detuned reflection coefficient  $\Gamma_d$  are derived.

The estimation of  $f_r$ , interpolating in between measured points, follows the evaluation of  $\Gamma_r$ , computed to be on the straight line connecting  $\Gamma_c$  and  $\Gamma_d$ :

$$\Gamma_{r,2} = 2\Gamma_c - \Gamma_d \quad (2.58)$$

Additionally, at each iteration, the new  $f_r$  serves to recompute the value of  $t$  and the  $n^{th}$  curve fitting procedure is performed to extract the coefficients,  $a_1$ ,  $a_2$  and  $a_3$  from  $\Gamma_{i,n}$ . Finally,  $\Gamma_c$ ,  $\Gamma_d$  and  $\Gamma_r$  are aligned with the origin of the Smith chart, as can be seen in Figure 2.7.

In the case of coupling losses, the resonance locus does not touch the circumference of the Smith chart. This means that the detuned reflection coefficient,  $\Gamma_d$  has an amplitude smaller than unity. This is attributed to the fact that the coupling is not entirely lossless. To account for it, it is common to add a series resistance in the equivalent model. In the Smith chart, an auxiliary circle of diameter  $D_2$  is introduced, which is tangential to both the resonance locus and Smith chart. This is used to realize the characterization of the resonator i.e. to derive the coupling coefficient from the diameter of the auxiliary circle and from the diameter of the resonance locus, as follows:

$$\kappa = \frac{1}{\frac{D_1}{D_2} - 1} \quad (2.59)$$

For the lossless coupling,  $D_2 = 2$  and (2.59) is identical to (2.55). From the curve-fitting procedure (2.57), the three coefficients  $a_1$ ,  $a_2$  and  $a_3$  are extracted and used to calculate  $D_2$  and  $Q_l$ , from which  $\kappa$ ,  $Q_0$  and  $Q_e$  are derived. The full calculation is described with more details in [Kajfez 94].

The output of the QZERO program is presented in Figure 2.7. Visible are 51 selected measurement points and the best fit of the resonance locus obtained from the QZERO software's iterations. The center of the resonance locus ( $\Gamma_c$ ) is indicated as well as the detuned reflection coefficient ( $\Gamma_d$ ), the loaded reflection coefficient ( $\Gamma_r$ ) corresponding to the loaded resonance frequency ( $f_r$ ). The values of the loaded Q-factor, "QL", unloaded Q-factor, "Q0", coupling coefficient, "K" and the interpolated loaded resonance frequency, "FL" are also indicated. The indication "LOSSY COUPLING" is indicative for the procedure used, either for the lossless or with loss coupling.



QZERO FOR WINDOWS (c) 2000 D. Kajfez

Data file name = B3

10/15/2010

16:14:

N= 51

LOSSY COUPLING

QL= 2895.3

+ - 28.8

Q0= 4514.2

+ - 67.2

K= 0.559135

+ - 0.01728

KS= 0.011258

FL= 4.00161664E+09

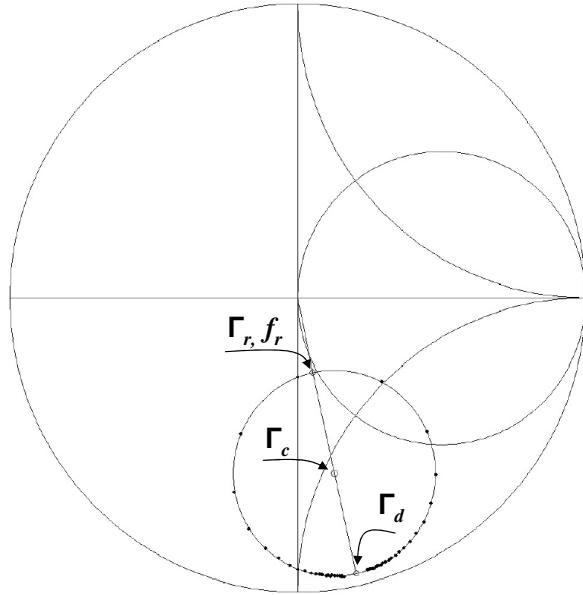


Fig. 2.7.: Output of the QZERO software [Kajfez 94]. Visible are the measurement points and the best fit circle. The center of the resonance locus ( $\Gamma_c$ ), the de-tuned reflection coefficient ( $\Gamma_d$ ), the loaded reflection coefficient ( $\Gamma_r$ ) and the loaded resonance frequency ( $f_r$ , "FL"). The values of the loaded Q-factor, "QL", unloaded Q-factor, "Q0" and coupling coefficient, "K". The indication "LOSSY COUPLING" is indicative for the procedure used.

## 2.4 PROTOTYPING AT 4 GHZ: FEEDING STRUCTURE ANALYSIS

In this section a rectangular cavity resonator is considered for 4 GHz, without any coupling. Next, three different coupling structures are investigated. Analytical developments are proposed and compared to full wave simulations with HFSS and measurements. The analysis of the first feed, a microstrip feeding through an aperture in the back side metallization is performed considering the energy transferred by the coupling structure to the resonator. The analysis of the second feed, a coplanar waveguide line (CPW) inserted for a given length in the cavity resonator is based on the variation of the input impedance with the location of the probing structure. The third feed, is a mix of the precedents with additional apertures provided at the location of the CPW short. Measurements are compared to full wave simulations and analytical developments.

### 2.4.1 The resonance frequency

A rectangular resonator is a rectangular waveguide with shorting plates, spaced by an integer number of wavelengths. The modes of resonance available in the resonator are both TE and TM. Each mode is defined by the subscripts  $m, n$  and  $l$  as  $TE_{mnl}$  and  $TM_{mnl}$ . The expression of the field distribution can be found in any reference book as [Montgomery 65, Collin 66, Pozar 05]. The characteristic field distribution of the fundamental  $TE_{101}$  mode of resonance is described in Figure 2.8.

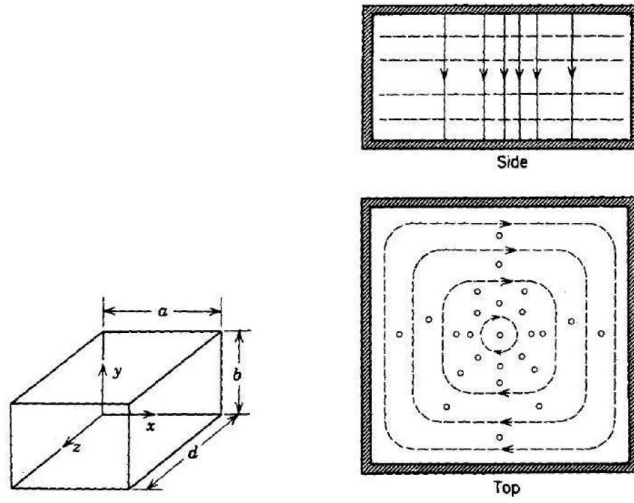


Fig. 2.8.: Rectangular cavity resonator, parameters definition and E field distribution of the fundamental  $TE_{101}$  mode of resonance [Gupta 79].

The  $TE_{101}$  field distributions in the cavity are given as follows:

$$E_y = E_0 \sin\left(\frac{\pi x}{a}\right) \sin\left(\frac{\pi z}{d}\right) \quad (2.60)$$

$$H_x = \frac{-j}{Z_{TE}} E_0 \sin\left(\frac{\pi x}{a}\right) \cos\left(\frac{\pi z}{d}\right) \quad (2.61)$$

$$H_z = \frac{j\pi}{k\eta a} E_0 \cos\left(\frac{\pi x}{a}\right) \sin\left(\frac{\pi z}{d}\right) \quad (2.62)$$

with  $Z_{TE} = k\eta/\beta_{01}$ , the wave impedance of TE propagation mode, where  $\eta = \eta_0/\sqrt{\epsilon_r}$  is the wave impedance,  $k = 2\pi f_{101}\sqrt{\epsilon\mu}$  and  $\beta_{01} = \sqrt{k^2 - (\pi/a)^2}$ , where  $\epsilon$  and  $\mu$  are the permeability and permittivity of the filling material. The electric field is perpendicular to the top and bottom surface at all points and magnetic field circulates around with no component in  $y$  direction. Since the electric field has only  $y$  components, opposite charges accumulate on opposite faces and a current flows through the side walls. Using

the expression of the fields of a traveling wave, propagating in the  $z$  direction through the waveguide, the propagation constant of the  $m, n^{th}$  TE or TM mode is given as:

$$\beta_{mnl}d = l\pi, \quad l = 1, 2, 3... \quad (2.63)$$

with  $\beta_{mn} = \sqrt{k^2 - \left(\frac{m\pi}{a}\right)^2 - \left(\frac{n\pi}{b}\right)^2}$ , the propagation constant of the  $m, n^{th}$  TE mode,  $b$  and  $a$  the height and width of the waveguide. The smallest possible cavity resonator encloses one half-guide wavelength, with  $l = 1$ . The wavenumber of the rectangular cavity is defined as [Pozar 05]:

$$k_{mnl} = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{l\pi}{d}\right)^2} \quad (2.64)$$

A direct link is established between the wavenumber  $k_{mnl}$  and the resonance frequency  $f_{mnl}$ , as follows:

$$f_{mnl} = \frac{k_{mnl}}{2\pi\sqrt{\mu\varepsilon}} = \frac{c_0}{2\pi\sqrt{\mu_r\varepsilon_r}} \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{l\pi}{d}\right)^2} \quad (2.65)$$

with  $c_0$  the speed of light,  $\mu_r$  and  $\varepsilon_r$  the relative permeability and relative permittivity of the material filling the cavity.

Associated is the unloaded Q-factor of the rectangular cavity resonator, with  $a \neq d$ , obtained considering the ratio between the electric or magnetic stored energy, calculated by integrating the fields over the cavity volume to the power loss, computed by integrating the loss per unit conductor area over the six cavity walls, given by [Pozar 05]:

$$Q_0 = \frac{\eta b}{2\pi^2 R_s} \left[ \frac{(kad)^3}{ad(a^2 + d^2) + 2b(a^3 + d^3)} \right] \quad (2.66)$$

with  $k = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{l\pi}{d}\right)^2}$ ,  $R_s = \frac{1}{\sigma\delta_s} = \sqrt{\frac{\omega\mu}{2\sigma}}$ , the surface resistance, where  $\sigma$  is the metal conductivity and  $\delta_s = \sqrt{\frac{2}{\omega\mu\sigma}}$ , the skin depth.

Finally, in the case of a square cavity resonator ( $a = d$ ) in the TE<sub>101</sub> resonance mode, the lateral dimensions are derived as follows:

$$a = \frac{\sqrt{2}\pi}{k_{mnl}} = \frac{c}{\sqrt{2}f_{mnl}} \quad (2.67)$$

and  $Q_0$  is rewritten as:

$$Q_0 = \frac{\pi\eta}{2R_s} \left[ \frac{\sqrt{2}b}{a + 2b} \right] \quad (2.68)$$

Following eigen mode full wave simulations, as well as [Harrington 01] increasing the height above half the lateral dimension, brings higher modes' resonance frequency ( $TE_{102}$  or  $TE_{201}$ ) closer to the fundamental mode ( $TE_{101}$ ). At 4 GHz, a rectangular cavity resonator is implemented with characteristic lateral dimensions:  $a = d = 53$  mm, maximizing the unloaded Q-factor. This cavity resonator is made in aluminum with specific conductivity  $\sigma_{Al} = 37.7 \cdot 10^6 \text{ Sm}^{-1}$  and therefore at this frequency  $R_s = 0.02 \Omega$ ,  $\delta_s = 1.3 \mu\text{m}$  and the expected unloaded Q-factor,  $Q_0 = 9946$  for a height,  $b = 25$  mm, which is lower than half the lateral dimensions, maximizing the isolation between adjacent resonance modes.

#### 2.4.2 Microstrip line feeding

A microstrip line coupling structure for a cavity resonator is thoroughly analyzed by Painchaud in the case of the coupling between a microstrip line and the whispering mode,  $TE_{102}$  mode of a cylindrical cavity resonator [Painchaud 76]. This approach is followed in this section, as it allows to analyze the feeding of a cavity resonator by a microstrip line.

The configuration of the  $TE_{101}$  square cavity resonator, under analysis and the characteristic parameters are presented in Figure 2.9-a). The cavity resonator at 4 GHz, mounted with the PCB and feed are presented in Figure 2.9-b). The microstrip line is located on top of the PCB and the aperture is opened in the PCB's back side metallization.

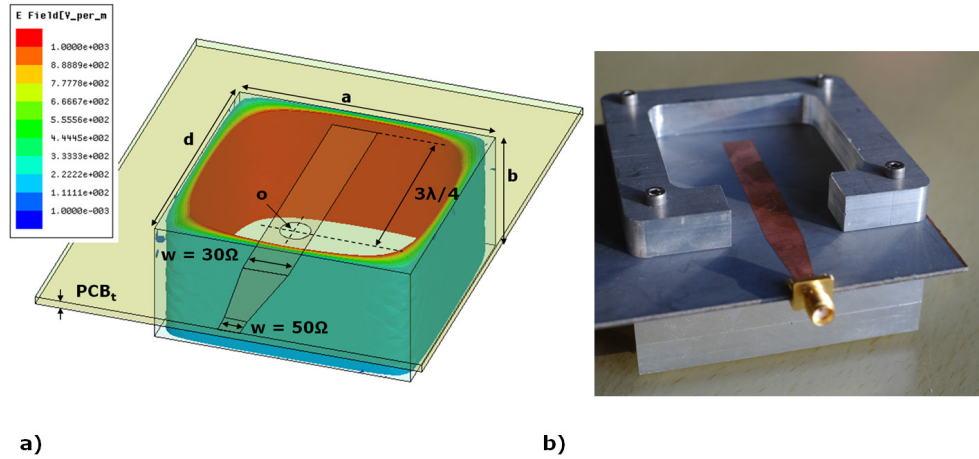


Fig. 2.9.: a) Microstrip feeding of the  $TE_{101}$ , E field distribution. b) Picture of the cavity resonator at 4 GHz mounted with feeding structure composed by the PCB with the microstrip line.

The coupling structure is composed by a microstrip line affected, at a certain distance from its end, by a circular aperture in the ground metallization. The distance from

the end is taken as such to maximize the magnetic coupling through the aperture, e.g. it has a length of three quarters of a wavelength in [Painchaud 76]. Instead, Harle is using a rectangular aperture located at a quarter of a wavelength from the end of the microstrip line [Harle 03]. This kind of coupling is also analyzed and re-named "*Feed From Above, FFA*" by Ocket [Ocket 09] in the case of the integration of a cavity resonator in MCM-D. His conclusions, following a parametric analysis are that the closer the slot position to the edge of the cavity, the higher the coupling; the higher the width of the rectangular aperture, the higher the coupling and that small variations of the stub length are negligible for the coupling values.

Following Painchaud, the analysis of the feed, as described in Figure 2.9 is based on the concept of transfer of energy, introduced by Wheeler for the analysis of the coupling between a waveguide and a cavity resonator [Wheeler 63]. In this approach, the coupling is evaluated by considering two symmetrical coupling problems. The first problem concerns the coupling between two waveguides due to the normalized waveguide reactance,  $x_{WG}$  of a small aperture in a common wall. The second problem, concerns the evaluation of the coupling coefficient,  $\kappa$  that exists between two identical cavities, coupled by an aperture in a common wall. The loading power between the cavity and the waveguide is the product of the normalized reactance by the coupling coefficient and is equal to the inverse of the external Q-factor,  $Q_e$ .

$$\frac{1}{Q_e} = \kappa x_{WG} \quad (2.69)$$

The equivalent energy concept, combined to a simplified parallel-plate waveguide model for the microstrip line is used to evaluate the magnetic coupling,  $\kappa_m$ , as follows [Painchaud 76]:

$$k_m = \frac{1}{4} \frac{V_{mc}}{V_m} \quad (2.70)$$

$V_{mc}$  is the effective volume of the coupling aperture and is evaluated following [Montgomery 65], for a circular aperture with diameter,  $o$ :

$$V_{mc} = \frac{2}{3} o^3 \quad (2.71)$$

In the idle case, in absence of the feeding, at the position of the aperture,  $V_m$  stores the same amount of energy as the full cavity.  $V_m$  is the effective volume of the cavity for magnetic coupling, satisfying the following relation [Painchaud 76]:

$$\frac{1}{2} \mu H_t V_m = \frac{1}{2} \mu \int_{cavity} H^2 dV \quad (2.72)$$

$V_m$  is isolated, as follows:

$$V_m = \frac{\int_{cavity} H_t^2 dV}{H_t^2} = \frac{W_m}{H_t^2} \quad (2.73)$$

where  $H_t$  is the tangential magnetic field at the location of the aperture but in absence of the aperture and the total magnetic energy stored in the  $TE_{101}$  mode in the cavity is given by [Pozar 05]:

$$W_m = \frac{\mu abd}{16} E_0^2 \quad (2.74)$$

Using the fields definition (2.60), the norm of the tangential magnetic field at the location of the aperture,  $H_t^2(x_0, z_0)$  can be directly evaluated. The effective volume is presented in Figure 2.10 as a function of the feed location. It can be observed that the effective volume is symmetrical with regard to the  $x$  and  $z$  axis, crossing in the center of the cavity resonator and is minimal at the middle of the lateral dimensions and as close as possible to the edge of the cavity. This observation motivates the choice to fix the location of the center of the circular aperture in the ground metallization at  $x_0 = 26.5$  mm and  $z_0 = 7.95$  mm, where the effective volume is among the lowest while not exactly on the edge, to avoid spurious perturbation. At this particular location, the ratio of the aperture in the ground metallization to the width of the feeding microstrip line is minimized.

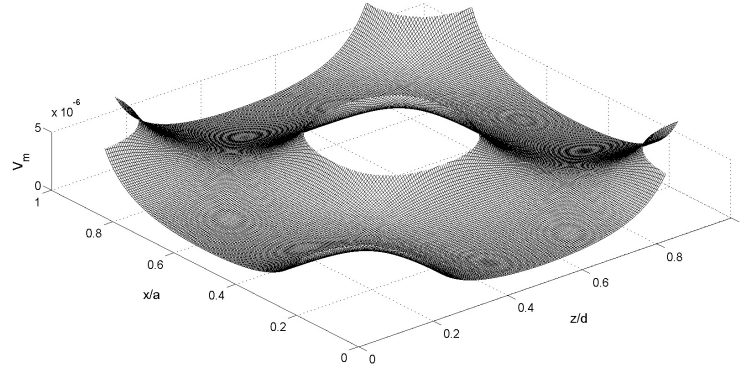


Fig. 2.10.: Effective volume of the  $TE_{101}$  square cavity resonator.

The analysis of  $x_{WG}$ , follows. The coupling between two identical microstrips is computed considering a parallel-plate waveguide model, as a simplified model for the microstrip. The value of the normalized impedance is given as follows [Painchaud 76]:

$$x_{WG} = \frac{\pi o^3}{3wh\lambda_g} \quad (2.75)$$

where  $w$  is the width of the microstrip line,  $o$  is the diameter of the aperture in between two waveguides,  $h_{PCB}$  is the thickness of the substrate and  $\lambda_g = \frac{\lambda_0}{\sqrt{\epsilon_r}}$  is the waveguide wavelength.

In our implementation using standard PCB material Duroid RT 5870 with a thickness:  $h_{PCB} = 1.524$  mm, relative permittivity:  $\epsilon_r = 2.33$  and dielectric losses, evaluated by:  $\tan \delta = 0.0012$ , the width of a microstrip line of  $50 \Omega$  is 4.66 mm and 9.42 mm for  $30 \Omega$ .

The external Q-factor (2.69) is evaluated using (2.70) and (2.75). Considering very large diameters are required, much larger than the microstrip's width for the  $50 \Omega$  microstrip line, the choice was made to use a lower impedance line, e.g.  $30 \Omega$  to reduce the diameter of the aperture. Finally, the coupling coefficient,  $\kappa$  is evaluated using (2.24), repeated here for reference:  $\kappa = Q_0/Q_e$  and considering the unloaded Q-factor defined in Section 2.4.1, formula (2.68).

A sampling set of five different resonators simulated with HFSS are compared to measurements and are presented in Figure 2.11. The values of coupling sweep from under critical through critical and overcoupled values as a function of the ratio  $o/w$ . The coupling coefficients and Q-factors extracted are given in Table 2.1. Following this comparison, it appears that they both are much lower in the measurements. This is due to the resistance of the melding between the metallized PCB and the cavity resonator, which is considered ideal in the HFSS simulations. Similarly, issues concerning the top and bottom cavity connections were reported by [Stickel 03].

Table 2.1.: Comparison between measured and simulated  $\kappa$  as a function of the ratio  $o/w$ . Unloaded Q-factors and normalization by  $\frac{Q_{0Meas.}}{Q_{0Sim.}}$ .

|       | $\kappa$   |           |       |                                       | $Q_0$        |             |
|-------|------------|-----------|-------|---------------------------------------|--------------|-------------|
| $o/w$ | Analytical | HFSS Sim. | Meas. | Sim. x $\frac{Q_{0Meas.}}{Q_{0Sim.}}$ | $Q_{0Meas.}$ | $Q_{0Sim.}$ |
| 0.50  | 0.245      | 0.232     | 0.062 | 0.044                                 | 1795         | 10482       |
| 0.55  | 0.434      | 0.443     | 0.092 | 0.084                                 | 1921         | 10581       |
| 0.60  | 0.731      | 0.684     | 0.117 | 0.130                                 | 1421         | 10558       |
| 0.65  | 1.182      | 1.058     | 0.364 | 0.201                                 | 2169         | 10535       |
| 0.70  | 1.844      | 1.590     | 0.609 | 0.303                                 | 2332         | 10500       |
| 0.75  | 2.789      | 2.299     | 1.152 | 0.437                                 | 2331         | 10470       |
| 0.80  | 4.109      | 3.119     | 1.752 | 0.593                                 | 2076         | 10415       |
| 0.90  | 8.329      | 5.870     | 1.488 | 1.117                                 | 1761         | 10318       |
| 1.00  | 15.673     | 8.941     | 2.329 | 1.701                                 | 1917         | 10151       |

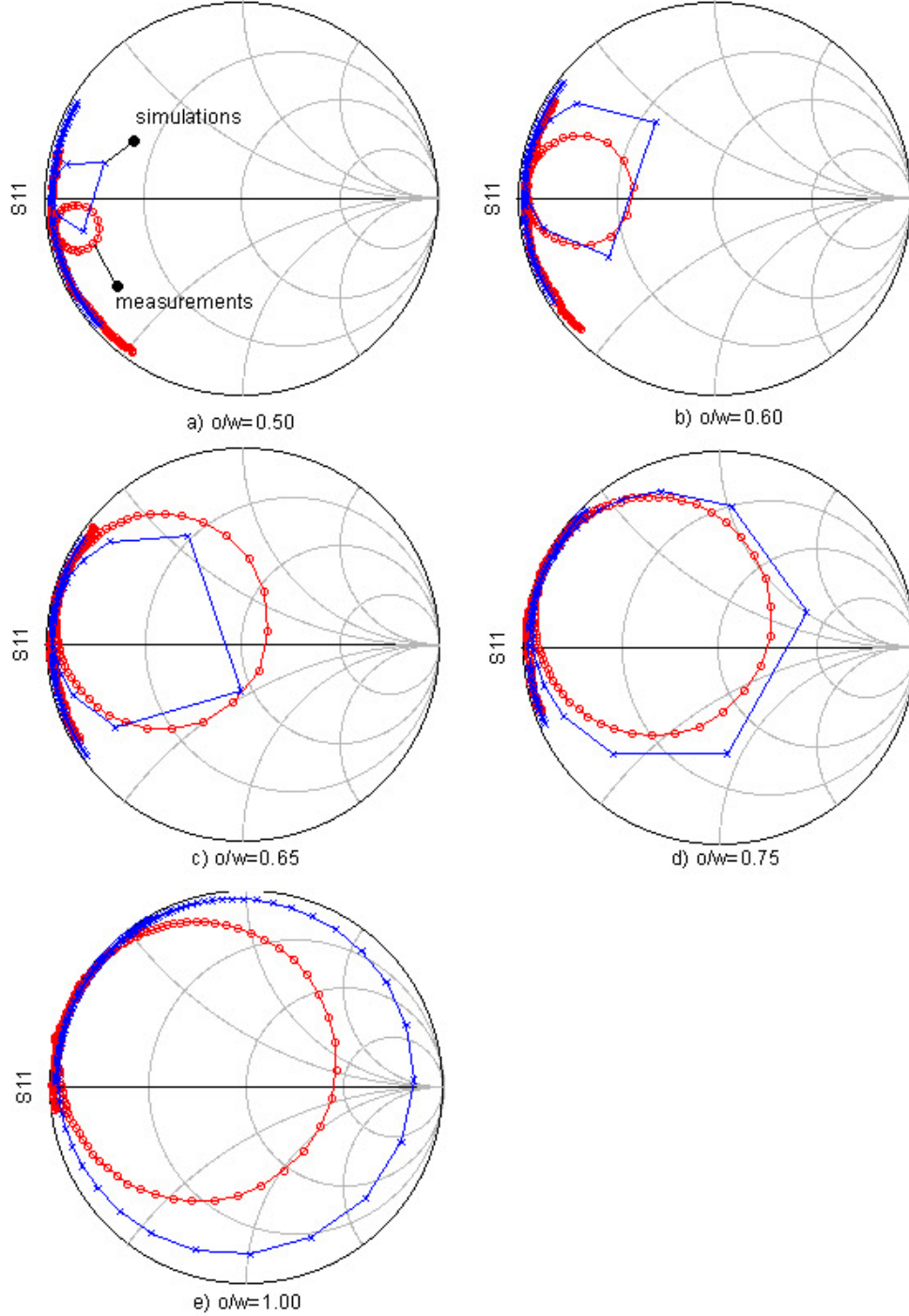


Fig. 2.11.: Comparison of the de-embedded HFSS simulations and measurements at 4 GHz of five different cavity resonators fed by a microstrip line with different microstrip feeding ratio aperture's opening diameter to the microstrip line's width, i.e.  $o/w = 0.5, 0.6, 0.65, 0.75$  and  $1$ , a) to e) respectively. Frequency range 3.7-4.3 GHz. De-embedding presented in Figure C.3, from Appendix C. Circle corresponds to measurements and  $\times$  corresponds to HFSS simulations.



Both HFSS simulations and measurements presented in Figure 2.11 were de-embedded using a Thru-Line-Reflection calibration (TLR) method described in [Eul 91]. This double de-embedding was required for accurate comparison, as for example the launching feature soldered to the microstrip line on PCB (coax to PCB) was not simulated in HFSS. The reference plane crosses the center of the circular aperture in the bottom PCB's metallization. The Reflection standard is here an Open, but could have been a Short, as well. The de-embedding steps are presented with more details in Figure C.3, from Appendix C. The straight lines in the Smith charts are due to the linear junction of measurement points.

To summarize, in Figure 2.12 are presented the analytical evaluation of the coupling factors, the extraction of the coupling coefficient from HFSS simulations and measurements of the  $TE_{101}$  mode of a rectangular cavity resonator coupled by a microstrip line, as a function of the ratio,  $o/w$  i.e. diameter of the aperture to the width of the microstrip line of  $30\ \Omega$ .

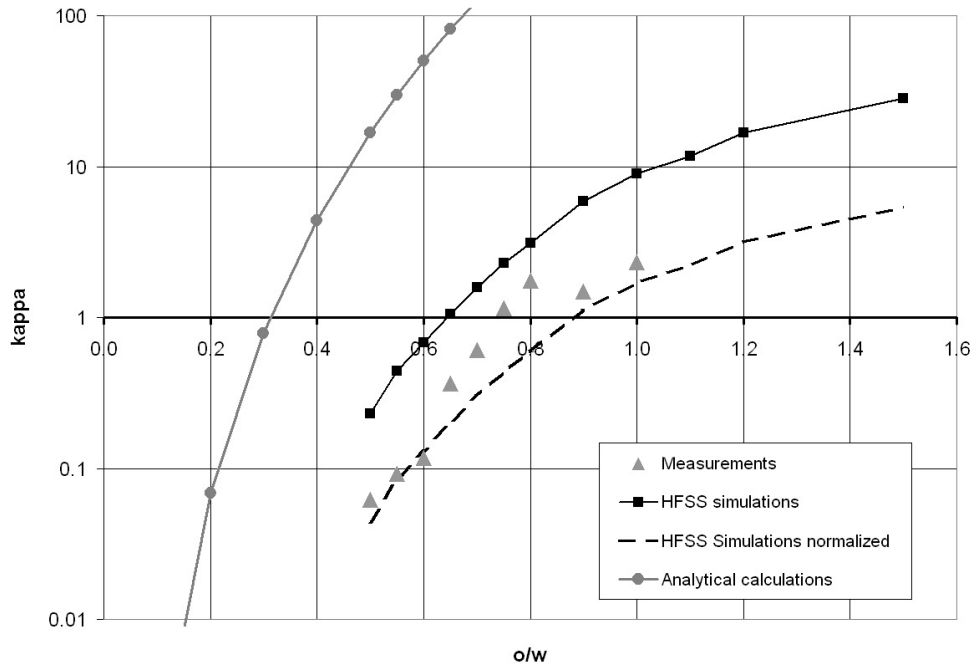


Fig. 2.12.: Comparison between the analytical evaluation, HFSS simulations and measurements of the coupling by a microstrip line of the  $TE_{101}$  mode of a rectangular cavity resonator.

In average, the extracted unloaded Q-factors from HFSS simulations is  $Q_{0av} = 10350$ , while the extracted Q-factors from the measurements, are in average  $Q_{0av} = 1969$ . The extracted Q-factors are presented in Table 2.1.

In what concerns the discrepancy between HFSS simulations and analytical evaluation, it is important to recall that in the analytical evaluation the fields distribution are considered not perturbed while this is not the case in reality. Additionally, in the HFSS simulations, additional sources of loss result from the mutual perturbation between the feeding structure and cavity resonator, increasing radiative losses and increasing reflection by changing the impedance. In what concerns the discrepancy between HFSS simulations and measurements, this is due to the additional resistive losses at the melding between the metallized PCB and the cavity resonator, which is considered ideal in the HFSS simulations. This is visible in the evaluation of the unloaded Q-factor and coupling coefficient. Actually, the coupling coefficient is proportional to the unloaded Q-factor. Considering the definition of the external Q-factor (2.24), repeated here for reference:  $\kappa = \frac{Q_0}{Q_e}$  we notice that a lower unloaded Q-factor,  $Q_0$  affects the coupling coefficient by decreasing it.

In order to take into account the additional sources of loss, predominately contact resistance, dielectric and radiative losses not enough accounted by the full wave simulations, we introduce the normalized coupling coefficient,  $\kappa_{norm.}$ . The normalized coupling coefficient is obtained by multiplying the simulated coupling coefficients with the ratio of Q-factors,  $\frac{Q_{0Meas.}}{Q_{0Sim.}}$ .

The ratio of Q-factors results from the evaluation of the measured to simulated unloaded Q-factors. As a consequence, the values of the normalized simulated coupling coefficients are getting closer to the measured values, as can be seen in Figure 2.12. This approach represents an alternative to the holistic method where all sources of loss are introduced into the simulation set. Actually, each of those sources are difficult to isolate and quantify, especially the above mentioned additional junction resistance.

$$\kappa_{norm.} = \kappa_{sim.} \frac{Q_{0Meas.}}{Q_{0Sim.}} \quad (2.76)$$

### 2.4.3 Coplanar waveguide feeding

A different approach is used to analyze the coupling of a cavity resonator by a Coplanar Waveguide (CPW) transmission line, patterned in the back side metallization closing the cavity resonator. This approach allows to consider multiple resonance modes and to calculate the impedance resulting at the feeding point. This analysis is usually used for patch antennas [Carver 81, Visser 05]. The antennas are considered as thin  $TM_z$ -mode cavity resonators, with magnetic walls. In the  $TM_z$ -mode as well as in the  $TE_{mnl}$ -modes, the electric field is perpendicular to the bottom, top plates. The field distribution is dual, because of dual boundary conditions on the lateral walls, respectively perfect magnetic (PMC) and perfect electric (PEC). The equations of Harrington for a coaxial feeding of a cavity resonator [Harrington 01] were adapted by Carver [Carver 81] to the case of a CPW feeding of a patch antenna, as described in Figure 2.13-a) and are revisited here in the case of a cavity resonator fed by a CPW line, as described in Figure 2.13-b).

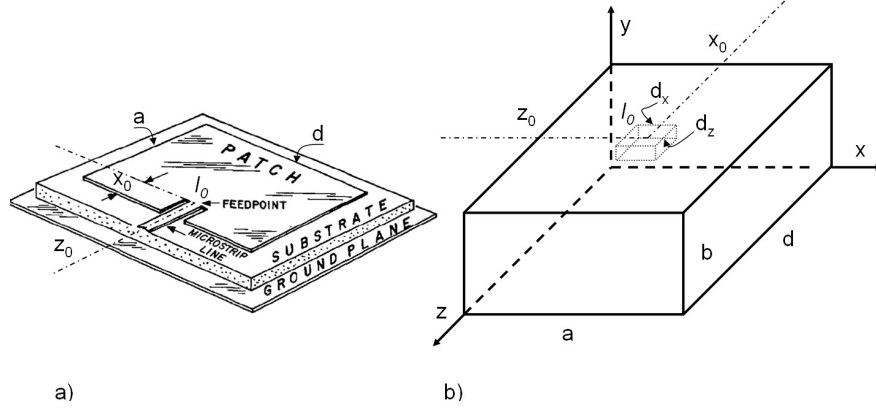


Fig. 2.13.: a) Coplanar Waveguide feed of the  $TM_z$  mode patch antenna [Carver 81]. b) The feed is represented by a rectangular equivalent current density  $I_0$  of lateral dimensions  $d_x$  and  $d_z$ , located at  $(x_0, z_0)$ , similarly in a  $TE_z$  cavity resonator.

The impedance at the input terminal of a probe inserted in a cavity resonator,  $Z_{in}$  is given as follows [Harrington 01]:

$$Z_{in} = \frac{1}{I_0^2} \int_V \vec{E} \cdot \vec{J}^* dV \quad (2.77)$$

where  $\vec{J}$  is the current distribution,  $\vec{E}$  is the field distribution of all modes,  $I$  is the corresponding input current and  $V$  corresponds to the inner volume of the cavity. As there are no losses introduced for the moment, the input impedance is purely imaginary. In order to solve (2.77), the fields are considered  $z$ -directed with only  $TE_{mn}$  modes available. The  $E_z$  field distribution is rewritten as [Carver 81]:

$$E_z = \sum_m \sum_n A_{mn} e_{mn} \quad (2.78)$$

where  $A_{mn}$  are the amplitudes of each mode and  $e_{mn}$  are the normalized  $E_z$  fields:

$$e_{mn} = \frac{1}{\sqrt{\varepsilon}b} \sin\left(\frac{m\pi x}{a}\right) \sin\left(\frac{n\pi z}{d}\right) \quad (2.79)$$

Considering a  $z$ -directed probe exciting the currents inside the cavity resonator by a current of magnitude  $I_0$ , with a small rectangular cross section  $(d_x, d_y)$  located at  $(x_0, y_0)$ , the magnitude of each mode is reduced to [Harrington 01]:

$$A_{mn} = jI_0 \sqrt{\mu b} \frac{k}{k^2 - k_{mn}^2} G_{mn} \sin\left(\frac{m\pi x_0}{a}\right) \sin\left(\frac{n\pi z_0}{d}\right) \quad (2.80)$$

where the normalized form factor,  $G_{mn}$  accounts for the excitation rectangular current element  $I_0$ , as can be seen in Figure 2.13-b) and is given by (2.81). In case of a CPW line feeding the resonator at  $(x_0, y_0)$ , one of the dimensions is set to the width of the transmission line while the other one is set to zero.

$$G_{mn} = \text{sinc}\left(\frac{n\pi d_x}{2a}\right) \text{sinc}\left(\frac{n\pi d_z}{2d}\right) \quad (2.81)$$

Considering:

$$\psi_{mn}(x, y) = \sin\left(\frac{m\pi x}{a}\right) \sin\left(\frac{n\pi z}{d}\right) \quad (2.82)$$

and substituting (2.80) and (2.79) into (2.78), we obtain:

$$E_z(x, z) = jI_0\eta \sum_m \sum_n \frac{k}{k^2 - k_{mn}^2} G_{mn} \psi_{mn}(x, z) \psi_{mn}(x_0, z_0) \quad (2.83)$$

where  $\eta = \sqrt{\frac{\mu}{\varepsilon}}$ ,  $k = \omega\sqrt{\mu\varepsilon}$  and  $k_{mn} = \sqrt{\frac{m\pi^2}{a^2} + \frac{n\pi^2}{d^2}}$

Finally, the voltage at the feed is computed as follows:

$$V_{in}(x_0, z_0) = -bE_z(x_0, z_0) \quad (2.84)$$

and the input impedance is given as [Carver 81]:

$$Z_{in}(x_0, z_0) = \frac{V_{in}}{I_0} = -j\eta kb \sum_m \sum_n \frac{\psi_{mn}^2(x_0, z_0)}{k^2 - k_{mn}^2} G_{mn} \quad (2.85)$$

If the cavity is lossy, but still maintaining a high Q-factor, the dissipation effect can be taken into account, by letting the eigen values to be complex by multiplying  $k_{mn}^2$  by  $(1 + \frac{j}{Q_{0mn}})$ , as following [Harrington 01].

In parallel with the analytical developments, full wave simulations are performed. The simulation set-up is described in Figure 2.14-a) and -b) where the parameters definitions are given and the E field distribution is plotted as resulting from the excitation by the given feed. Similarly to the microstrip feeding, standard PCB material is used (Duroid RT 5870:  $h_{PCB} = 1.524$  mm,  $\varepsilon_r = 2.33$  and  $\tan \delta = 0.0012$ ). As an illustration, the width of the 50  $\Omega$  Coplanar Waveguide line is  $CPW_w = 2.16$  mm and the slot  $CPW_s = 100$   $\mu m$ . In this configuration, an opening above the CPW line in the metallic casing needs to be provided. Following parametric simulations with HFSS, it results that an opening with a length of 8 mm and a height of 4 mm is negligibly affecting the RF characteristics.

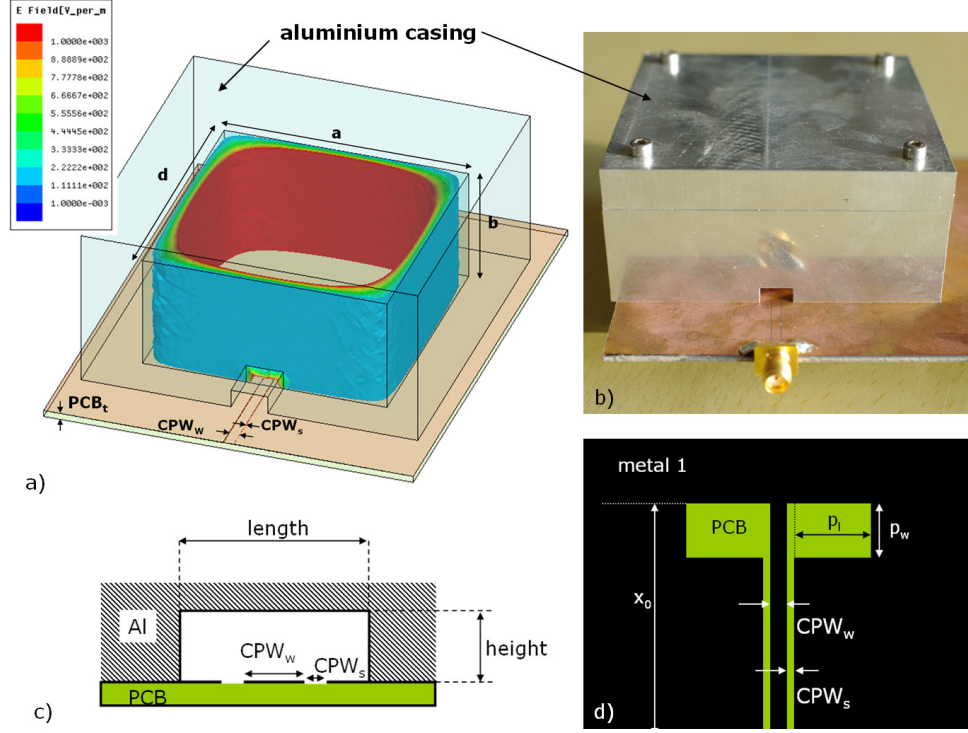


Fig. 2.14.: a) Parameters of the Coplanar Waveguide feeding of the  $TE_{101}$  mode of resonance of the 4 GHz cavity resonator and E field distribution. b) Realized aluminum casing prototype. c) Cross section of the opening in the aluminum casing. d) Parameters of the aperture located at the end of the CPW line.

The extracted coupling factors and corresponding Q-factors are given in Table 2.2. A sampling set of five different resonators simulated with HFSS and measured are presented in Figure 2.15. With the aim of extrapolating the results at other frequencies, the results are given as a function of the ratio  $x_0/a$ , where  $x_0$  is the location of the short of the CPW line on the x-axis, starting from the edge of the cavity ( $x_0 = 0$ ), as indicated in Figure 2.14 and  $a$  is the corresponding lateral dimension of the cavity resonator. The values of coupling sweep from under critical through critical and overcoupled values as a function of the ratio  $x_0/a$ . Corresponding extracted coupling values and Q-factors are given in Table 2.1. The measurements were performed on the cavity resonator described in Figure 2.14-b), where the CPW line enters in the cavity through the opening in the metallic casing. As previously mentioned, both HFSS simulations and measurements were de-embedded using a TLR method [Eul 91]. For comparison with HFSS simulations, after de-embedding, the reference plane is located at the short end of the CPW line, provided by  $x_0/a$ . More details are presented in Figure C.5, from Appendix C. The straight lines in the Smith charts are due to the linear junction of measurement points.

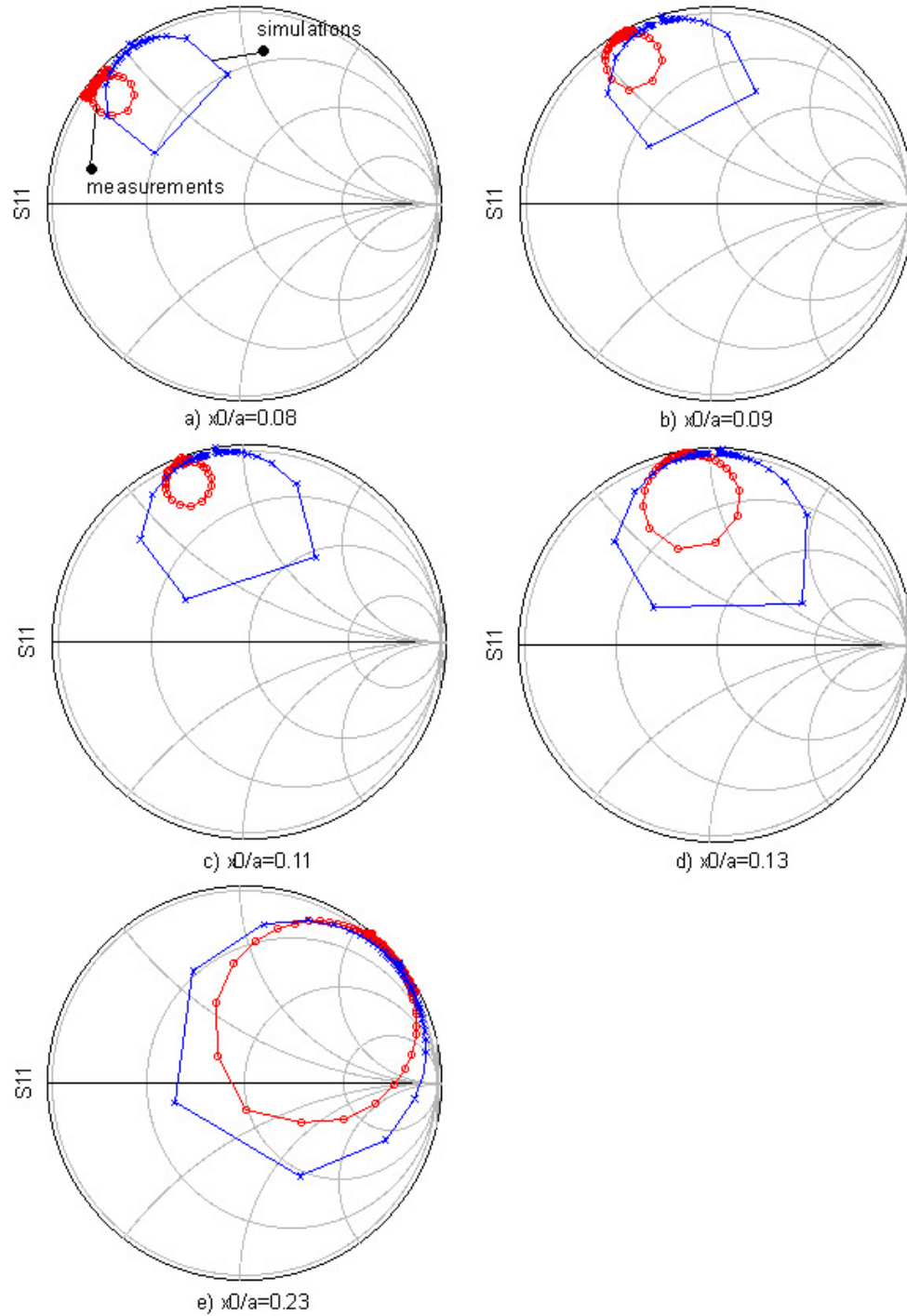


Fig. 2.15.: Comparison of de-embedded HFSS simulations and measurements of the 4 GHz cavity resonators with shorted CPW feed, located at different positions, starting from inside the cavity, at its edge, i.e.  $x_0/a = 0$  and for  $x_0/a = 0.08, 0.09, 0.11, 0.13$  and  $0.23$ . De-embedding steps presented in Figure C.5, from Appendix C. Circle corresponds to measurements and  $\times$  corresponds to HFSS simulations.

In Figure 2.16 are presented the analytical evaluation of the coupling factors, the extraction of the coupling coefficient from HFSS simulations and measurements of the rectangular cavity resonator coupled by a CPW line, as a function of the ratio:  $x_0/a$ . The same conclusions stated previously for the microstrip line feed, concerning the discrepancy between full wave simulations, analytical evaluation and measurements, hold. The lower values of coupling are due to lower unloaded Q-factors measured. In average, the extracted unloaded Q-factors from HFSS simulations is  $Q_{0av} = 10351$ , while the extracted Q-factors from the measurements, are in average  $Q_{0av} = 2972$ . The coupling coefficients measured are also approximatively three times smaller than simulated with HFSS, as we can see from Table 2.2.

Table 2.2.: Comparison between measurements and simulated coupling coefficient,  $\kappa$  as a function of the ratio  $x_0/a$ . Unloaded Q-factors and normalization by  $\frac{Q_{0Meas.}}{Q_{0Sim.}}$ .

|         | $\kappa$   |           |       |                                       | $Q_0$        |             |
|---------|------------|-----------|-------|---------------------------------------|--------------|-------------|
| $x_0/a$ | Analytical | HFSS Sim. | Meas. | Sim. x $\frac{Q_{0Meas.}}{Q_{0Sim.}}$ | $Q_{0Meas.}$ | $Q_{0Sim.}$ |
| 0.08    | 1.371      | 0.537     | 0.103 | 0.169                                 | 2634         | 10390       |
| 0.09    | 1.726      | 0.635     | 0.147 | 0.200                                 | 3135         | 10363       |
| 0.11    | 2.544      | 0.985     | 0.197 | 0.310                                 | 2658         | 10351       |
| 0.13    | 3.497      | 1.071     | 0.304 | 0.337                                 | 3254         | 10313       |
| 0.15    | 4.569      | 1.646     | 0.501 | 0.517                                 | 3180         | 10322       |
| 0.17    | 5.745      | 1.590     | 0.624 | 0.500                                 | 3018         | 10291       |
| 0.23    | 9.696      | 2.048     | 1.102 | 0.677                                 | 3183         | 10276       |

The importance of the characterization by measurements, as emphasized by [Kajfez 94] is perfectly understood. Actually, Ohmic losses, as those of the additional junction resistance, are drastically reducing the Q-factors and values three times lower than expected are commonly measured. Additionally, the only opening in the back side metallization is the slot of the CPW line ( $CPW_s = 100 \mu m$ ). This is reducing the magnetic coupling as the field lines are cut. Additionally, issues are common as such dimensions are critical for low complexity photolithography and wet etch PCB manufacture. Therefore, the normalization by the ratio of Q-factors is very interesting when accurate values of coupling and impedance are required.

The symmetry observed in Figure 2.16 for the analytical development is as a consequence of the symmetry of the field distribution. Even though the accuracy of the analytical development is not sufficient e.g. for co-design in an oscillator, it gives a tremendous insight of the behavior of this particular kind of feeding.

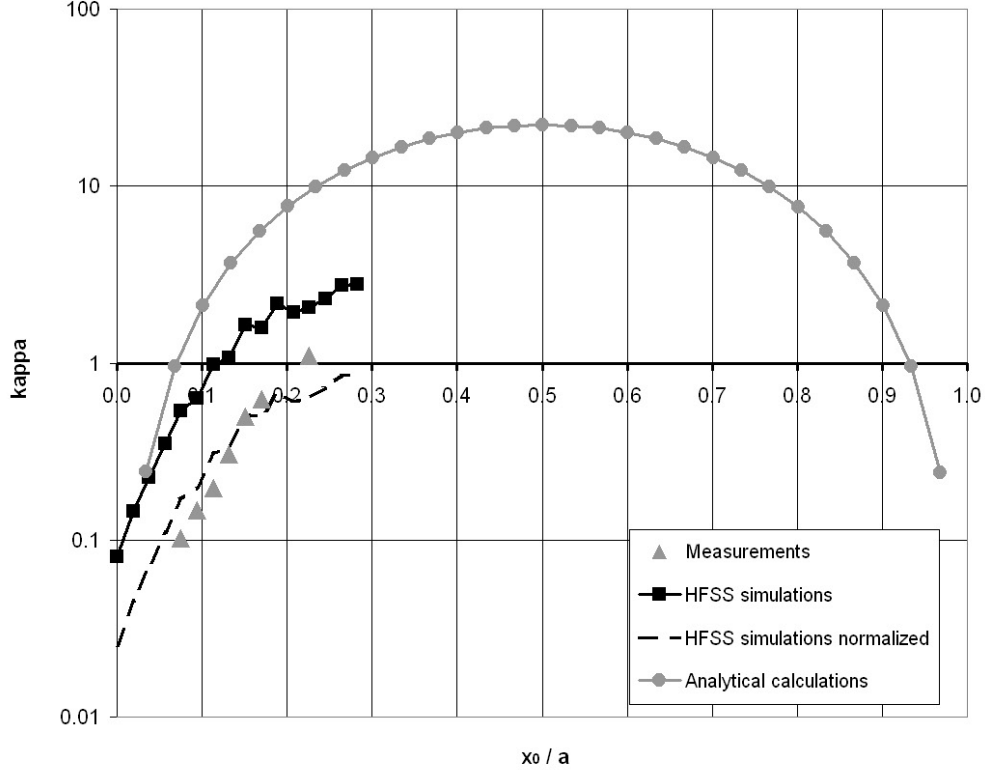


Fig. 2.16.: Comparison between the analytical evaluation, HFSS simulations and measurements of the coupling by a CPW line of a  $TE_{101}$  rectangular cavity resonator.

#### 2.4.4 Coplanar waveguide feeding ended with additional apertures

Similarly to the feeding using a shorted Coplanar Waveguide, shorted inside the cavity resonator at a distance  $x_0$  from the edge, additional apertures are provided at the shorted end, as defined in Figure 2.14-d). The parameters defining this additional aperture are the width,  $p_w$  and the length,  $p_l$  of the aperture in the back metallization. The length,  $p_l$  is taken after the CPW's slot ( $CPW_s = 100 \mu m$ ). At the location of the short, the magnetic field around the shorted CPW line is maximum. The aperture un-shields the field lines, the larger the aperture the larger the coupling coefficient. The analysis of the feed is performed starting from the location where the critical coupling was obtained for the shorted CPW feed ( $x_0/a = 0.13$ ), see Section 2.4.3.

A sampling set of five resonators simulated with HFSS and measured is presented in Figure 2.17. Both HFSS simulations and measurements are de-embedded using the TLR method described in [Eul 91]. The de-embedding steps are presented in Figure C.6, from Appendix C. The straight lines in the Smith charts are due to the linear junction of measurement points.



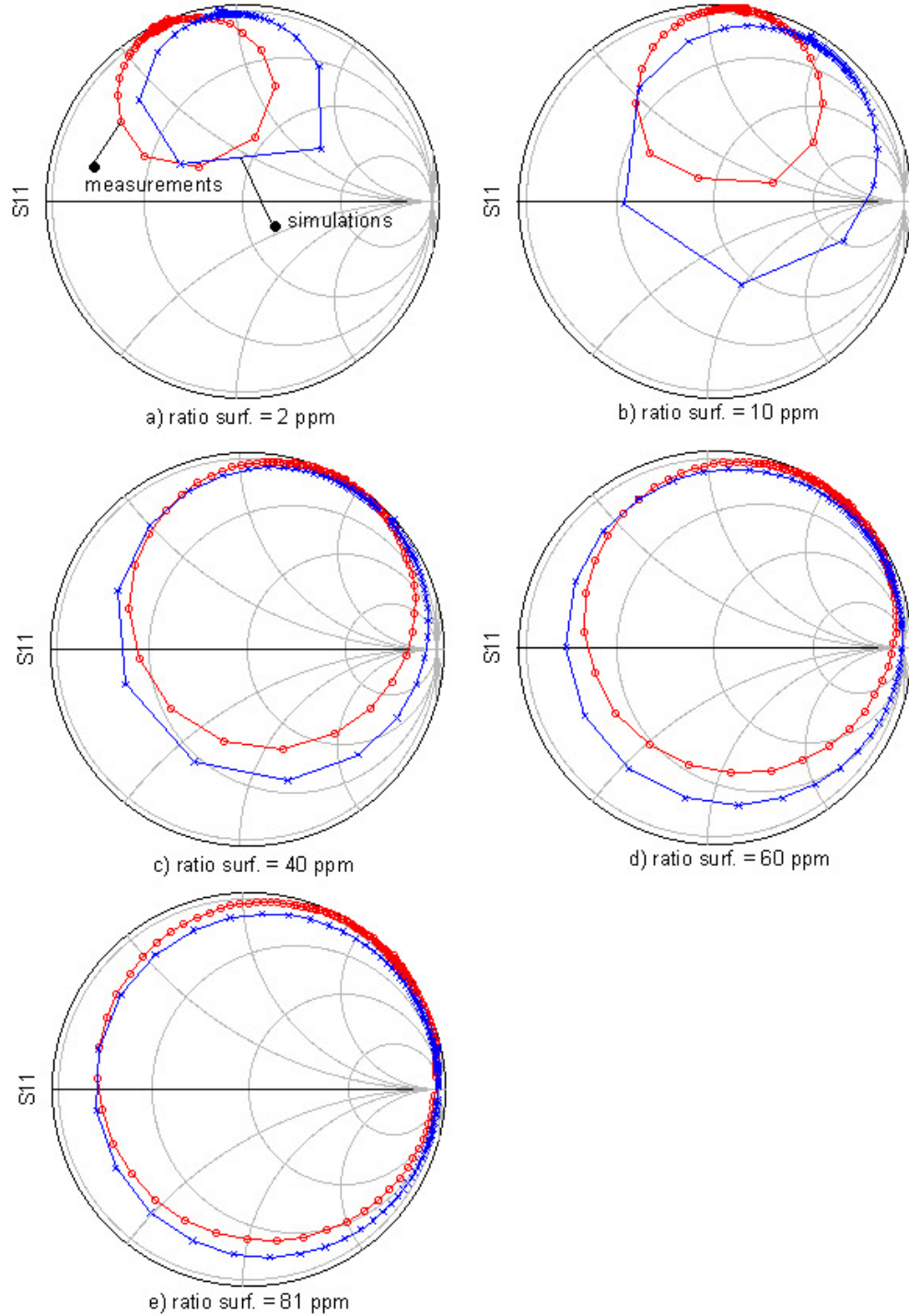


Fig. 2.17.: Comparison between the de-embedded HFSS simulations and measurements at 4 GHz of the cavity resonators with CPW feed provided with additional ending aperture of different ratios surf. (aperture area to the bottom cavity's area): 2, 10, 40, 60 and 81 ppm. The ratio  $x_0/a = 0.13$  is used for all variations. De-embedding steps are presented in Figure C.6, from Appendix C. Circle corresponds to measurements and  $\times$  corresponds to HFSS simulations.

The set starts with  $p_l = 0$  which is equivalent to the case when the aperture is the slot of the CPW line ( $CPW_s = 100 \mu m$ ). The width of the aperture is maintained constant,  $p_w = 3 \text{ mm}$  while  $p_l$  is swept from 0 to 5 mm. In order to remain generic and to scale the results for other frequencies, the coupling coefficient is given as a function of the ratio of areas between the aperture's surface and the cavity's area:

$$\text{ratio surf.} = \frac{p_w p_l}{ad} \quad (2.86)$$

where  $p_w$  and  $p_l$  are respectively the width and length of the aperture in the PCB's back metallization.

The corresponding extracted coupling values are given in Table 2.3 where are also provided the extracted Q-factors. The overview comparison between measurements, simulations and normalized simulations is shown in Figure 2.18.

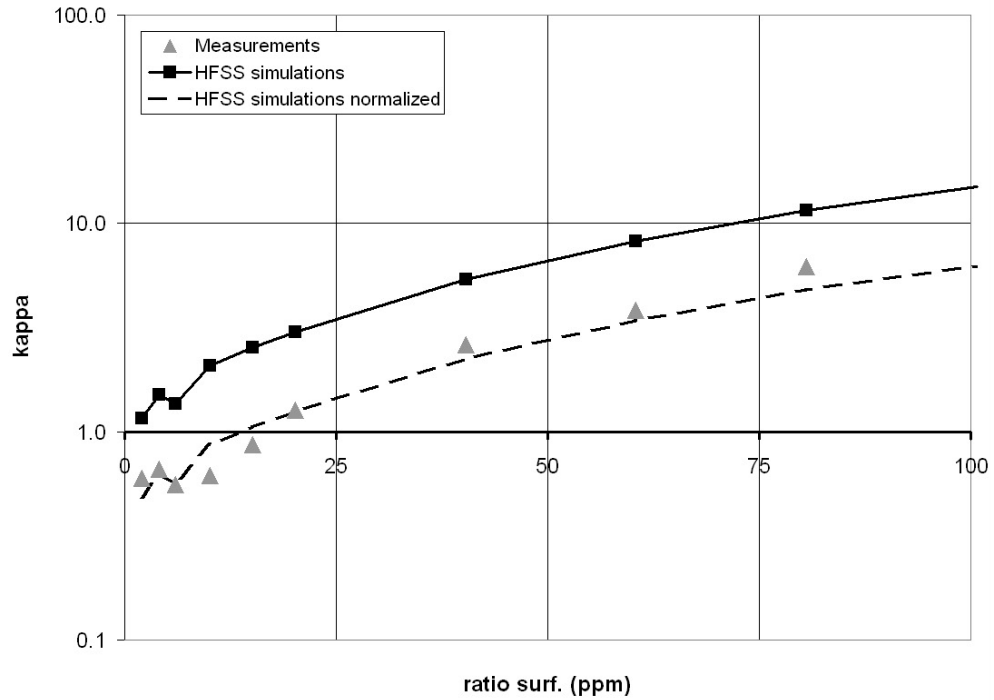


Fig. 2.18.: Comparison between the HFSS simulations, measurements and normalized simulations of the coupling of the rectangular cavity resonator by a CPW line ended with an aperture. The coupling coefficient is given as a function of the areas ratio between the slot opening and the area of the cavity.

Table 2.3.: Comparison between measurements and HFSS simulations as a function of ratio surf. Unloaded Q-factors and normalization by  $\frac{Q_{0_{Meas.}}}{Q_{0_{Sim.}}}$ .

|             | $\kappa$  |       |                                             | $Q_0$           |                |
|-------------|-----------|-------|---------------------------------------------|-----------------|----------------|
| ratio surf. | HFSS Sim. | Meas. | Sim. x $\frac{Q_{0_{Meas.}}}{Q_{0_{Sim.}}}$ | $Q_{0_{Meas.}}$ | $Q_{0_{Sim.}}$ |
| 2 ppm       | 1.160     | 0.599 | 0.484                                       | 4331            | 10370          |
| 10 ppm      | 2.074     | 0.616 | 0.865                                       | 3974            | 10328          |
| 40 ppm      | 5.376     | 2.626 | 2.243                                       | 4274            | 10040          |
| 60 ppm      | 8.229     | 3.814 | 3.432                                       | 3755            | 9770           |
| 81 ppm      | 11.515    | 6.221 | 4.803                                       | 3892            | 9321           |

Actually, the input impedance is a function of both the location of the feed and of the power transferred to the resonator through the aperture. The analytical analysis is not performed here, instead full wave simulations are compared to measurements.

The average of the extracted Q-factor from simulation is  $Q_{0av} = 9980$  which can be compared to the average extracted Q-factor from measurement,  $Q_{0av} = 4163$ . The simulations normalized by the ratio of Q-factors are in very good agreement with the measurements, as can be seen Figure 2.18.

#### 2.4.5 Discussion and recommendations

*A priori*, and following HFSS simulations, all three feeds are similar: the shorted CPW feed is ranked first followed by the microstrip line feed and finally by the shorted CPW line, with additional apertures. The average unloaded Q-factors are respectively  $Q_{0av} = 10351$ ,  $Q_{0av} = 10350$  and  $Q_{0av} = 9980$ , as can be seen in Figure 2.19. The Q-factors are extracted with QZERO. For comparison purposes only, the Q-factors are not given as a function of their main parameter but indicatively as per simulation. Following the HFSS simulations, the Q-factors are higher than theoretically expected from a full aluminum based cavity resonator ( $Q_0 = 9946$ ). This is due to the back side metallization which is made of copper and contributes to decreasing the Ohmic losses. Any opening in the ground plane metallization is decreasing the unloaded Q-factor which is more pronounced in the case of the shorted CPW feed with additional apertures and the larger the aperture the lower the Q-factor. The unloaded Q-factors are also decreasing with larger openings in the ground plane metallization, in the case of the microstrip line feed. The larger the ratio  $o/w$ , the larger the radiative losses. The less affected is the shorted CPW, where only the slot of the CPW line ( $100 \mu m$ ) is opening the ground plane metallization.

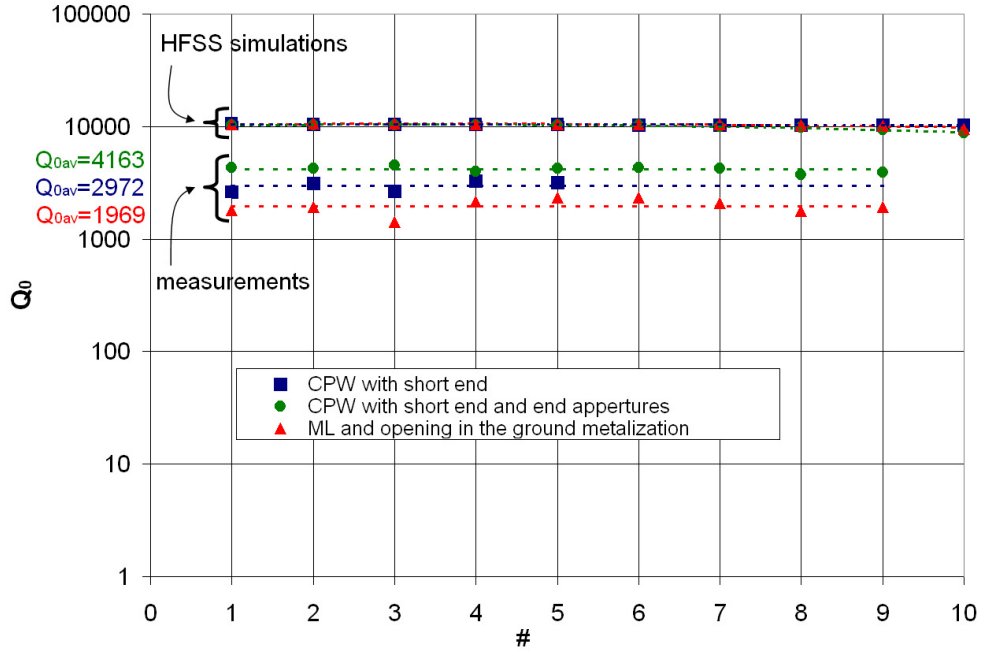


Fig. 2.19.: Comparison between the HFSS simulations and measurements of different couplings.

*A posteriori*, and following the measurement results, the shorted CPW line feed, with additional apertures is ranked first, followed by the shorted CPW line feed and finally by the microstrip line feed, as can be seen in Figure 2.19.

The highest measured unloaded Q-factor is  $Q_0 = 4330$  for a shorted CPW line feed with a ratio surf. of 2 ppm. The larger the aperture the lower the Q-factor. On average, for the shorted CPW with additional apertures, the unloaded Q-factor is  $Q_{0av} = 4163$ . The average unloaded Q-factor of the shorted CPW is:  $Q_{0av} = 2972$ . Finally the microstrip line feed has a measured unloaded Q-factor, on average  $Q_{0av} = 1969$ .

The difference in the ranking between measurements and simulations results essentially from the additional resistive losses at the junction between the metallized PCB and the cavity resonator, which is considered ideal in the HFSS simulations. Additionally, higher radiative, Ohmic and dielectric losses could also be accounted and dielectric losses of the PCB affect the unloaded Q-factor by decreasing it.

The discrepancy observed between simulation and measurements can be corrected in a holistic approach, by adapting the simulation's set-up to the measured characteristics of the cavity resonator: the dielectric substrate, the conductivity of the aluminum casing, introduction of an equivalent lossy material mimicking the junction of the copper metallization to the aluminium cavity and radiation losses. The normalization of the Q-factors extracted from simulations (2.76) by the ratio of the average Q-factors:

$\frac{Q_{0Meas.}}{Q_{0Sim.}}$ , is our recommended methodology. Actually, the losses parameters are difficult to isolate, especially the junction and contact losses but also radiation losses.

## 2.5 60 AND 75 GHZ SUBSTRATE INTEGRATED CYLINDRICAL CAVITY RESONATOR

After having investigated the theory of cavity resonators, assessed the characterization methodology and developed different feeding structures in order to understand the coupling mechanism, in this section we design a cylindrical Silicon integrated Cavity Resonator (SiCR) for 60 and 75 GHz. Due to the high ratio values  $D/H_{HR-Si} \approx 10$  (diameter to the height of the SiCR,  $D = 1130 \mu m / H_{HR-Si} = 100 \mu m$ ) the  $TM_{010}$  field distribution of the cylindrical cavity resonator is very similar to the  $TE_{101}$  field distribution of the square cavity resonator, while the Q-factor is slightly improved. The SiCR is fully integrated in IMEC's MCM-D integration platform [Dancila 11a]. We introduce in Figure 2.20 the new topology, defined by a double row of TSVs aligned with a pitch of  $200 \mu m$ , imposed by the technology and referring to [Posada 08].

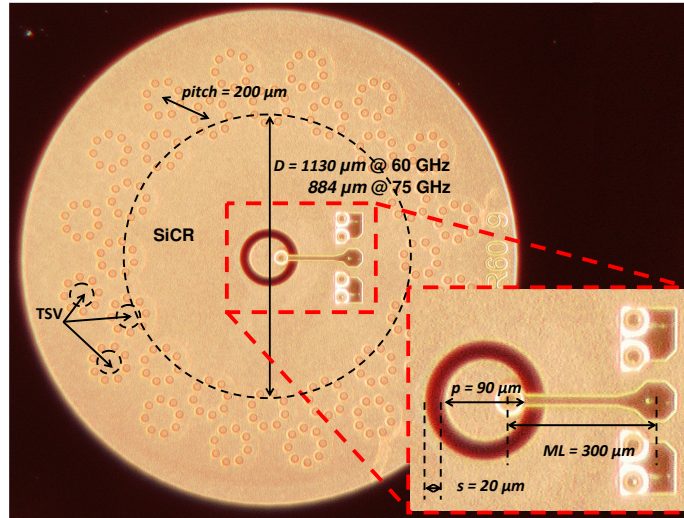


Fig. 2.20.: Silicon integrated cavity resonator, SiCR. Visible are the double rows of TSVs, the new feed composed by a metallic patch of diameter,  $p$  and the connecting microstrip line, provided with probing pads.

Adapted to the technology, a new feeding structure is proposed. This is composed of a metallic patch which diameter,  $p$ , see Figure 2.20, fixes the coupling coefficient. A theoretical and experimental characterization is conducted, the following parameters are extracted: the unloaded resonance frequency, the unloaded Q-factor and the coupling coefficient.

The extraction of the unloaded resonance frequency is realized in Section 2.6 and is based on an empirical method, following [Canos 06]. Based on this characterization a novel method of extraction the lumped elements equivalent circuit is derived in Section 2.7.

The main advantages of the SiCR proposed are first the very good performance, typical unloaded Q-factors,  $Q_0$  of 117 and 91 at respectively 60 and 75 GHz. Actually, integrated inductors for low cost (Bi)CMOS or Bipolar technologies are typically characterized by a low Q-factor, limited to around 30 below 10 GHz and decreasing with increasing operation frequency [Tilmans 03]. The performance improvement results from the fully shielded implementation which reduces radiation losses, spurious couplings and substrate excitations at very high frequencies. Secondly, the realization of the substrate integrated cavity resonators do not require additional manufacturing given the availability of the interconnection TSVs, as would be required by the capping of the air cavity resonator. Furthermore, the level of compactness and integration is improved through the high permittivity of the filling material: HR-Si, is characterized by a high permittivity which reduces the overall size of the resonators by a factor 3.4 in comparison with the air cavities. Finally, the metallic encapsulation of the resonator allows to re-use the top side shielding as landing surface for e.g. flip chipping of active devices, MMICs, entering in an oscillator design, as presented in [Dancila 11b]. This increases further the level of integration improving the multi-chip module concept.

### 2.5.1 Technology

The cross section of the MCM-D integration platform with integrated passives and cavity resonator is shown in Figure 2.21. IMEC's MCM-D platform is largely described by [Posada 07] and [Carchon 01]. The platform, incorporates integrated passives realized in the top thin metallization layers produced by using sputter methods. The metallizations are embedded in benzo-cyclobutene (BCB)  $\varepsilon_{BCB} = 2.65$ . The first layer, 1  $\mu m$  Al is sputtered directly on top of the HR-Si and covered by 4.4  $\mu m$  BCB. The second layer, 5  $\mu m$  Cu is covered by 8.5  $\mu m$  BCB. The third layer is an electroplated stack of 10  $\mu m$  Cu/Ni/Au.

The thinning of the thick wafers and processing of the TSVs starts after the patterning of the metallizations and dielectric layers. The MCM-D wafers are bonded faced down with thick wax to a carrier wafer using wafer level bonding wafer and thinned down to 100  $\mu m$  by mechanical grinding using both coarse and fine grinding. The total thickness variation (TTV) over a wafer was below 5  $\mu m$ . The bonded wafers are cleaned to remove particles; and the surface damage after grinding is removed by dry etch step. Through the 100  $\mu m$  thinned HR-Si wafer, TSVs are processed in a three step etching. The first 30  $\mu m$  are isotropically etched, followed by an anisotropic etch using a Bosch etch process [Robert Bosch GmbH]. Additionally, a soft landing process is used to obtain a smoother surface at the junction with the top side metallization. During this process the plasma energy is lowered before reaching the substrate. Following the three steps processing, the TSVs are shaped in the particular form presented in Figure 2.22. The TSVs are characterized by a bottom diameter of 100  $\mu m$  and a top diameter of 70  $\mu m$ . The details about the optimization of this etch have been presented in [Majeed 09].



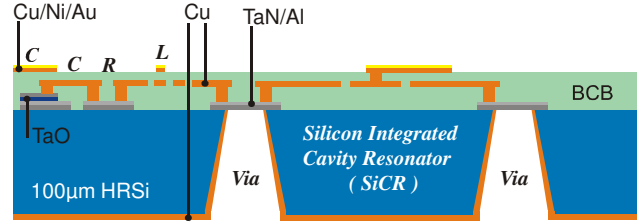


Fig. 2.21.: Cross section of the thin-film MCM-D, 100  $\mu\text{m}$  thinned HR-Si substrate with SiCR shaped by TSVs. The MMIC chip is directly flip-chipped on landing structures patterned in the third metallization.

Finally, the TSV are conformally plated with 5  $\mu\text{m}$  Cu, as well as the back side of the wafer with a two step processing. First, a low-current density is applied to have a gradual deposition to enhance the Cu seed layer on the sidewalls. Second, a higher current density is used for bulk of the Cu plating. In the last step of processing the negative photoresist is removed and the seed layer is removed with a wet etch chemistry.

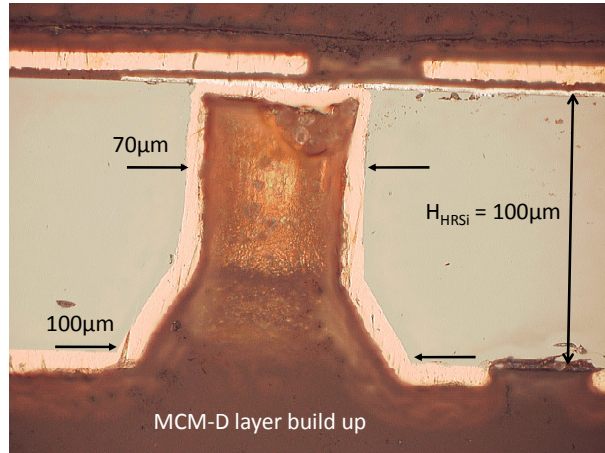


Fig. 2.22.: Through silicon vias, TSV. The diameter of the upper part of the TSV is 70  $\mu\text{m}$  and of the lower part is 100  $\mu\text{m}$ .

### 2.5.2 The resonance frequency

The circular shape of the cavity resonator was preferred to the square one as this improves slightly the unloaded Q-factor, respectively  $Q_{0_{circular}} = 170$  vs.  $Q_{0_{square}} = 166$ , as following eigen mode simulations realized at 60 GHz with Ansoft's HFSS [HFSS] and considering both BCB and HR-Si filling materials, as described in Figure 2.21. The cylindrical cavity resonator is a section of circular waveguide of length  $H_{HR-Si}$  and diameter  $D$ , with short circuiting plates at each end, as defined in Figure 2.23.

The possible modes are the TE and TM forms [Montgomery 65]. Inside such a resonator, an infinity of linear superposition of modes is present. Each mode is defined by the subscripts  $l, n$  and  $m$  as  $TE_{lm}$  and  $TM_{lm}$ . The field distributions can be found in any reference books as [Collin 66, Pozar 05]. Here the  $TM_{010}$  mode is the dominant mode for the very large ratios:  $D/H_{HR-Si} \approx 10$ .

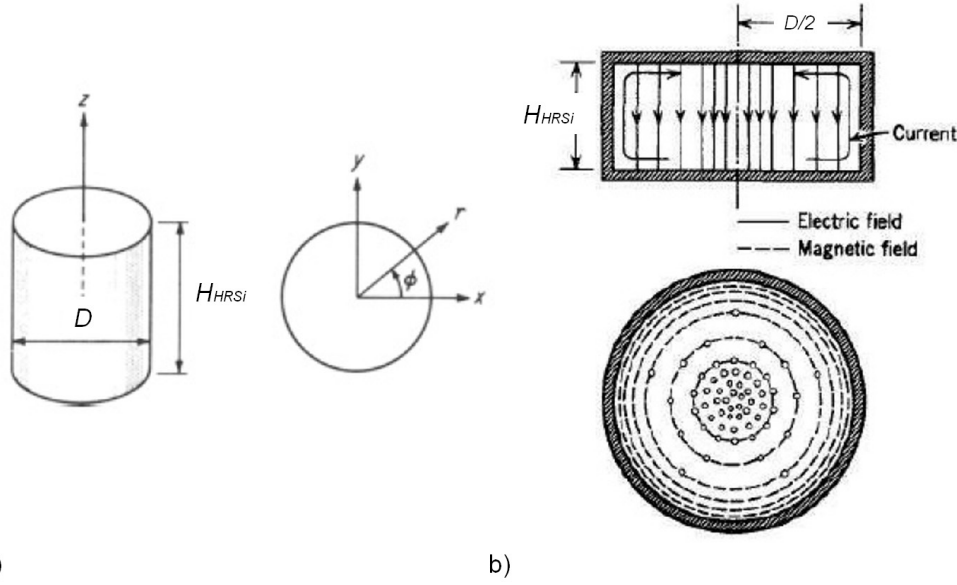


Fig. 2.23.: a) Cylindrical cavity resonator, parameters definition and b)  $TM_{010}$  electric and magnetic fields distribution [Gupta 79].

The fields distribution for the  $TM_{010}$  mode is given as follows:

$$E_z = E_0 J_0(kr) \quad (2.87)$$

$$H_\phi = E_0 \frac{j}{\eta} J_1(kr) \quad (2.88)$$

with  $k = \frac{p_{01}}{a} = \frac{2.405}{a}$ , the resonance frequency,  $f_0$  is given by [Pozar 05]:

$$f_0 = \frac{c_0}{\pi \sqrt{\epsilon_r} D} \quad (2.89)$$

with  $c_0$  the speed of light in vacuum. Equation (2.89) corresponds to the fundamental mode of resonance  $TM_{010}$ , of a cylindrical cavity resonator.

Following the technology description in Section 2.5.1, the relative dielectric permittivity  $\epsilon_r$  is evaluated considering the series stacking of two dielectric layers filling the



cavity resonator, as follows:

$$\varepsilon_r = \frac{\varepsilon_{BCB}\varepsilon_{HR-Si}(d_{BCB} + d_{HR-Si})}{(\varepsilon_{BCB}d_{HR-Si} + \varepsilon_{HR-Si}d_{BCB})} \quad (2.90)$$

Using (2.90) the relative permittivity is  $\varepsilon_r = 10.4$ , considering the stacking formed by  $4.4 \mu m$  BCB and  $100 \mu m$  HR-Si.

The resonance frequency, as a function of the diameter of the resonator is calculated using (2.89) and (2.90) and is presented in Figure 2.24, Figure 2.25 and Table 2.4.

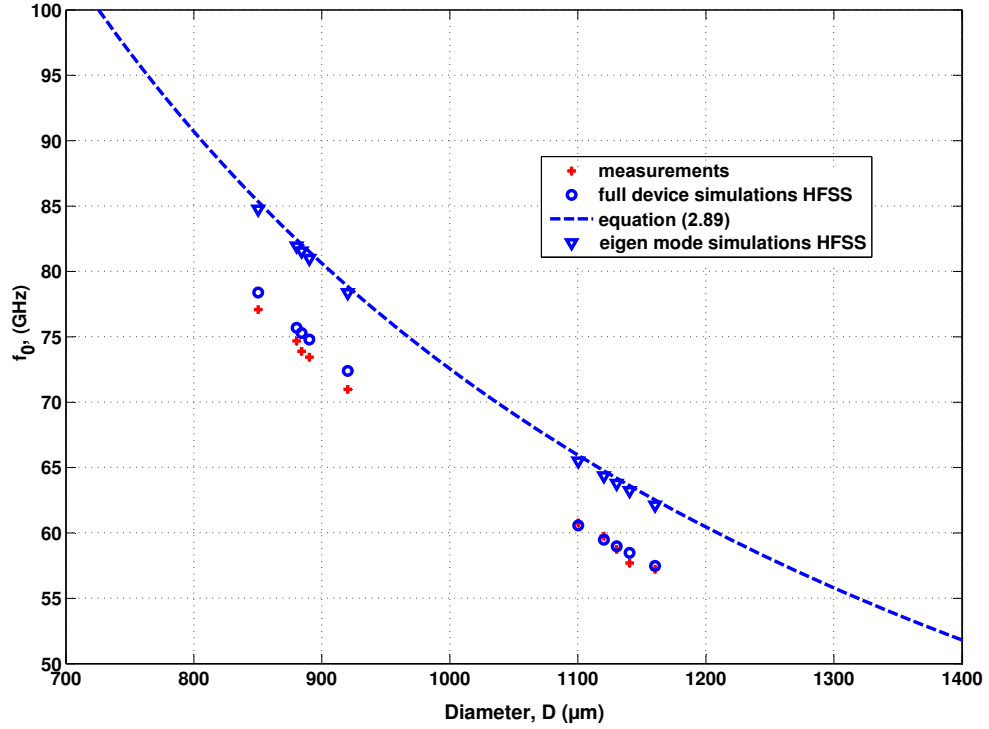


Fig. 2.24.: Measured and simulated influence of the cavity resonator's diameter,  $D$  on the resonance frequency. The diameter of the resonators are indicated in Table 2.4. The diameter of the feeding patch is  $p = 80 \mu m$ .

The analytical calculation using (2.89) and eigen mode simulations realized with HFSS, both considering a cylindrical resonator with diameter  $D$  are compared to measurements and full wave simulations performed with HFSS, that consider the actual topology with TSVs located tangential to the cylinder used for analytical and eigen mode formulations.

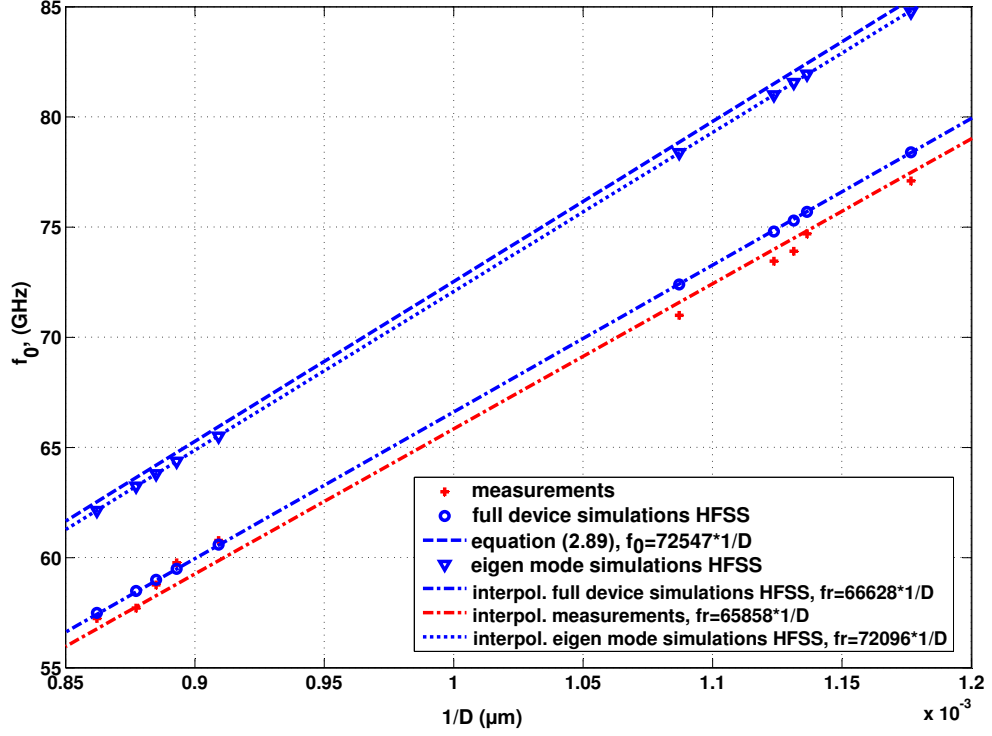


Fig. 2.25.: Extrapolated data from measured, the full-wave HFSS simulations, the eigen mode simulations and the analytical expression, (2.89) of the resonance frequency,  $f_0$  as a function of the diameter,  $D$ . Resonators' diameter are indicated in Table 2.4. The diameter of the feeding patch is  $p = 80 \mu\text{m}$ .

In Figure 2.25, the linear interpolations illustrate the  $1/D$  dependency of the resonance frequency and compare easily the measurements with HFSS simulations, eigen mode simulations with HFSS and the analytical expression (2.89). We conclude, as described in Table 2.4 that the evaluation by the eigen mode simulations with HFSS are lower by 0.6% on average than (2.89). The HFSS simulations provide a lower resonance frequency than evaluated analytically (2.89) and by eigen mode with HFSS. When compared to the measurements, the full-wave HFSS simulations are on average 1.2% lower, because the effective diameter is larger. Finally, to link measurements and HFSS simulation to analytical evaluations (2.89) and eigen mode HFSS simulations, we introduce the effective diameter,  $D_{eff}$ :

$$D_{eff} = D/\eta_{eff} \quad (2.91)$$

where  $\eta_{eff}$  is the effective shielding coefficient, evaluated here to be  $\eta_{eff} = 0.9884$ , following the linear interpolations presented in Figure 2.25. The shielding coefficient,  $\eta_{eff} < 1$  is meaningful of fringing fields between adjacents TSVs. The resonance fre-

quency measured is lower than simulated and therefore the effective diameter enclosed by the TSVs is larger. For correction the diameter  $D_{eff}$  should be considered in (2.89).

Table 2.4.: Comparison between measured and simulated frequency response computed with HFSS considering the actual topology with TSVs located tangential to the cylinder used for analytical and eigen mode calculation, given as a function of the diameter of the resonator.

| $D$<br>( $\mu m$ ) | $p$<br>( $\mu m$ ) | Sim. HFSS<br>$f_r$ (GHz) | Meas.<br>$f_r$ (GHz) | $\Delta$<br>(%) |
|--------------------|--------------------|--------------------------|----------------------|-----------------|
| 1100               | 80                 | 60.6                     | 60.79                | 0.31            |
| 1120               | 80                 | 59.5                     | 59.76                | 0.40            |
| 1130               | 80                 | 59.0                     | 58.76                | -0.43           |
| 1140               | 80                 | 58.5                     | 57.70                | -1.34           |
| 1160               | 80                 | 57.5                     | 57.23                | -0.48           |
| 850                | 80                 | 78.4                     | 77.00                | -1.74           |
| 880                | 80                 | 75.7                     | 74.13                | -1.95           |
| 884                | 80                 | 75.3                     | 73.92                | -1.86           |
| 890                | 80                 | 74.8                     | 73.47                | -1.85           |
| 920                | 80                 | 72.4                     | 71.00                | -1.97           |

With an increasing operation frequency, the diameter of the cylindrical cavity resonator is reducing. Given the technology used and the minimal pitch available, limitations will appear concerning the operation of the cavity resonator. The diameter of the cavity resonator is imposed by the pitch and the diameter of the TSV: the pitch length diminished by a TSV's diameter. For the smallest diameter available, the openings in between the TSVs are large and would not allow the energy to be stored without high radiative losses. Usually, the openings in between two TSVs are advised to be in the order of  $\lambda/20$  to  $\lambda/10$ . Additionally, more than one row of TSVs could be used to shape the cavity resonator, with the expense of an increased surface. Nevertheless, this could reduce the risk of failure due to the TSVs yield and is increasing the unloaded Q-factor, by reducing the radiative losses, as demonstrated in [Posada 08].

In Figure 2.26 and in Figure 2.27 we present three samples of the reflection coefficient measurements ( $S_{11}$  parameter), summarily presented in Figure 2.24 and Figure 2.25. Resonators' diameter are indicated in Table 2.4. The nominal diameter of the feeding patch is  $p = 80 \mu m$ , at both frequencies, while the resonators' diameter is respectively  $D = 1130 \mu m$  and  $D = 884 \mu m$ .

At 60 GHz, Figure 2.26, the extracted Q-factors from measurements are on average  $Q_0 = 117.1$  and  $Q_l = 81.4$ . In comparison the extracted Q-factors from the HFSS simulations, are on average  $Q_0 = 163.4$  and  $Q_l = 97.2$ .

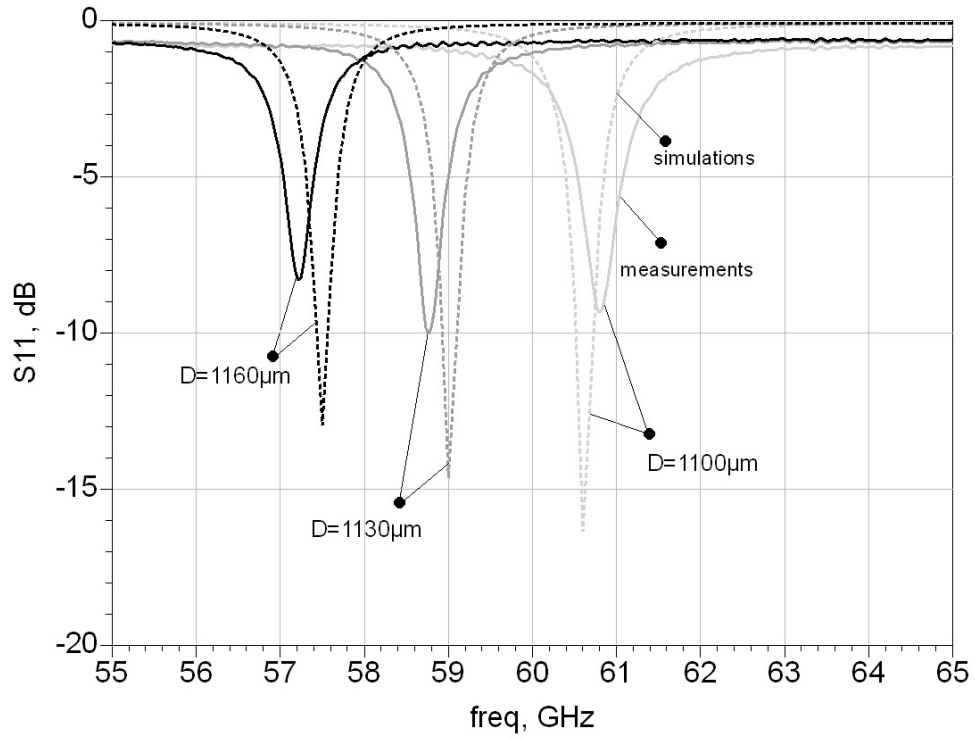


Fig. 2.26.: Measured (symbol line) and simulated (solid line) influence of the cavity resonator's diameter,  $D$  on the resonance frequency ( $S_{11}$  parameter) around 60 GHz. Resonators' diameter are indicated in Table 2.4. The nominal diameter of the feeding patch is  $p = 80 \mu m$ .

At 75 GHz, Figure 2.27, the extracted Q-factors from measurements are on average  $Q_0 = 91.2$  and  $Q_l = 52.2$ . In comparison the extracted Q-factors from the HFSS simulations, are on average  $Q_0 = 174.4$  and  $Q_l = 69.8$ .

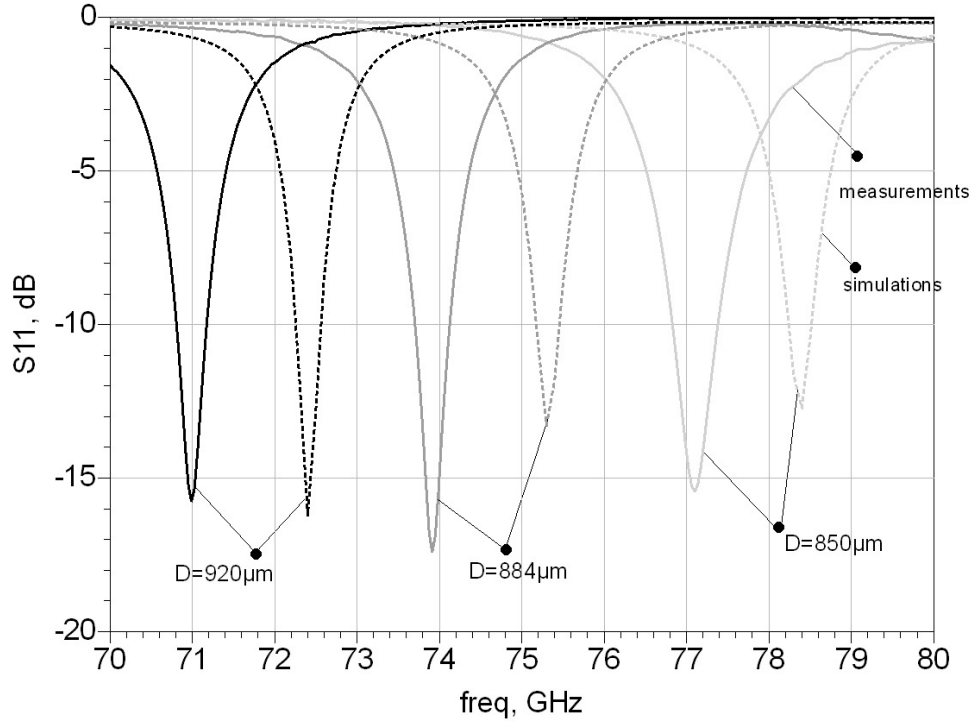


Fig. 2.27.: Measured (symbol line) and simulated (solid line) influence of the cavity resonator's diameter,  $D$  on the resonance frequency ( $S_{11}$  parameter) around 75 GHz. Resonators' diameter are indicated in Table 2.4. The nominal diameter of the feeding patch is  $p = 80 \mu\text{m}$ .

### 2.5.3 Coupling coefficient and Q-factors extraction

The new patch feeding introduced in Figure 2.20 and in Figure 2.21 is composed of a microstrip line, realized in the third metallization, connects through small vias, through the BCB, a circular patch patterned in the second metallization. Given the BCB thickness of  $8.5 \mu\text{m}$  in between the metallizations the width of the  $50 \Omega$  microstrip line is  $20 \mu\text{m}$ . This coupling is well adapted to the available MCM-D technology while offering variable and largely available values of coupling,  $\kappa$  by changing the patch diameter,  $p$ .

The patch feeding couples electrically to the resonator's field distribution and excites the  $\text{TM}_{010}$  fundamental mode of the cylindrical resonator. This coupling is visible in Figure 2.28, where the electric field distribution inside the cavity resonator is highlighted. The fields are attenuated by more than 40 dB after the first row of TSVs. This is in good agreement with [Posada 08], where a double row of TSVs with the same pitch of  $200 \mu\text{m}$  and a diameter of  $100 \mu\text{m}$  demonstrates a very good shielding at 60 GHz.

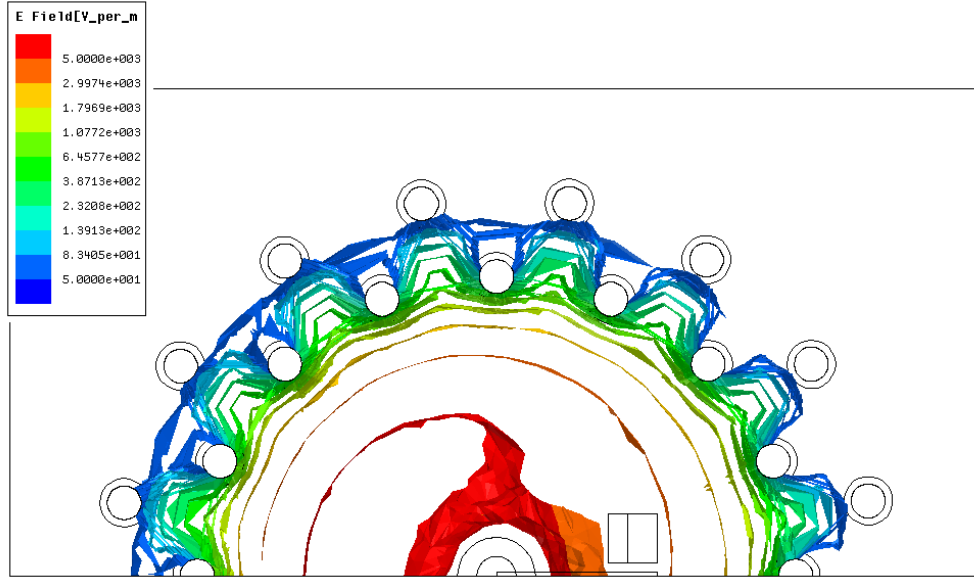


Fig. 2.28.: E field distribution of the excitation of the fundamental  $TM_{010}$  mode of the cylindrical cavity resonator, coupled by the proposed patch feeding.

The substrate integrated cavity resonators implement different diameters of the feeding patch,  $p$  to obtain different impedances and couplings coefficients,  $\kappa$ . The values of  $\kappa$  sweep from undercoupled values ( $\kappa < 1$ ) to overcoupled values ( $\kappa > 1$ ). Keeping the resonator's diameter constant ( $D = 1130 \mu m$ ), the variation of the coupling coefficient is realized by changing the patch diameter,  $p$  as defined in Figure 2.20. These resonators are simulated with HFSS in Figure 2.29, measured (Figure 2.30), and the simulations and measurements are compared in Figure 2.31 and in Table 2.5.

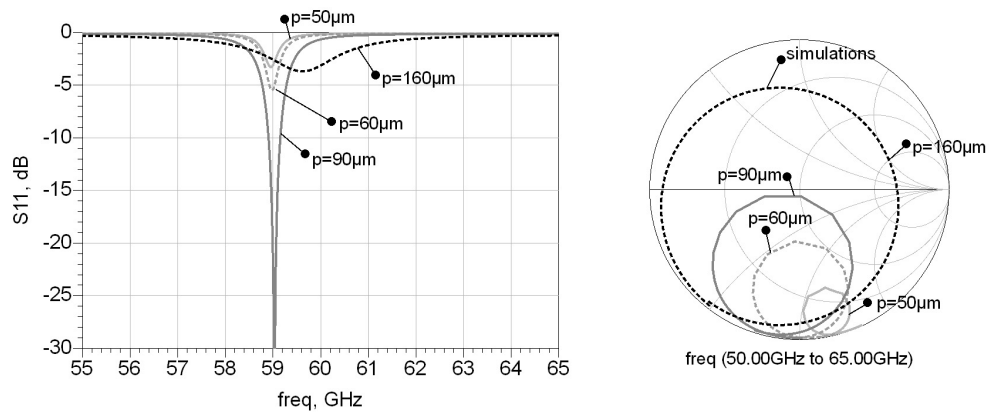


Fig. 2.29.: HFSS simulations of the reflection coefficient,  $S_{11}$  parameter as a function of the diameter,  $p$  ( $D = 1130 \mu m$ ).

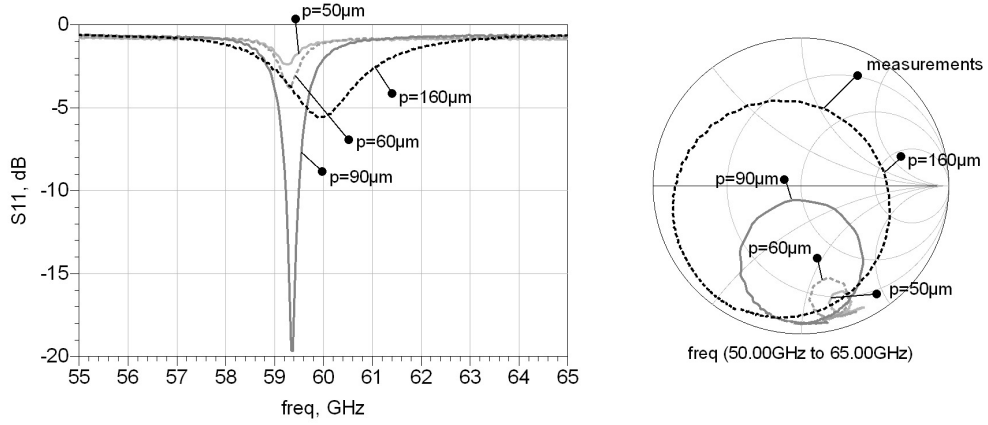


Fig. 2.30.: Measured reflection coefficient,  $S_{11}$  parameter.

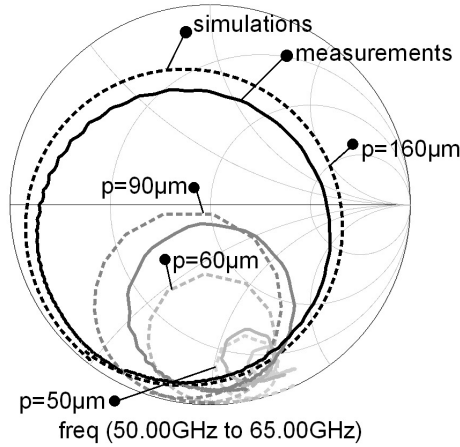


Fig. 2.31.: Comparison between measured and simulated influence of the diameter  $p$  of the feeding patch on coupling.

The characteristics of any cavity resonator need to be defined by measurements [Kajfez 94] and are here extracted with QZERO. Those characteristics are the unloaded resonance frequency,  $f_0$  the unloaded Q-factor,  $Q_0$  and coupling coefficient,  $\kappa$ . They are considered following the definitions developed in Section 2.3.

The  $Q$ -factors extracted from HFSS simulations and measurements are presented in Table 2.5 and in Table 2.6. In both cases, the variation of the coupling coefficient as a function of  $p$  is given in Figure 2.32. The coupling coefficient monotonically increases with an increasing  $p$ . The critical coupling occurs for  $p = 90\mu\text{m}$  and  $p = 70\mu\text{m}$  at respectively 60 and 75 GHz for the HFSS simulations and for  $p = 110\mu\text{m}$  and  $p = 95\mu\text{m}$  at respectively 60 and 75 GHz for the measurement. In this case,  $\kappa = 1$  and an equal amount of power is dissipated in the external circuit as in the resonator,

therefore  $Q_e = Q_0$  and  $Q_l = Q_0/2$ . Apparently, at higher frequencies, a smaller patch diameter ensures the same level of coupling. Meanwhile the diameter of the cavity resonator is also smaller i.e.  $1130$  vs  $884 \mu m$  and the ratio  $p/D$  is kept constant. The discrepancy between HFSS simulations and measurements is higher at  $75$  GHz, resulting from the fact that the substrate losses are not variable with the frequency and are kept constant for all frequencies. Obviously, the losses are more important at higher frequencies, which is affecting the coupling coefficient ( $\kappa$  is decreased), as previously mentioned.

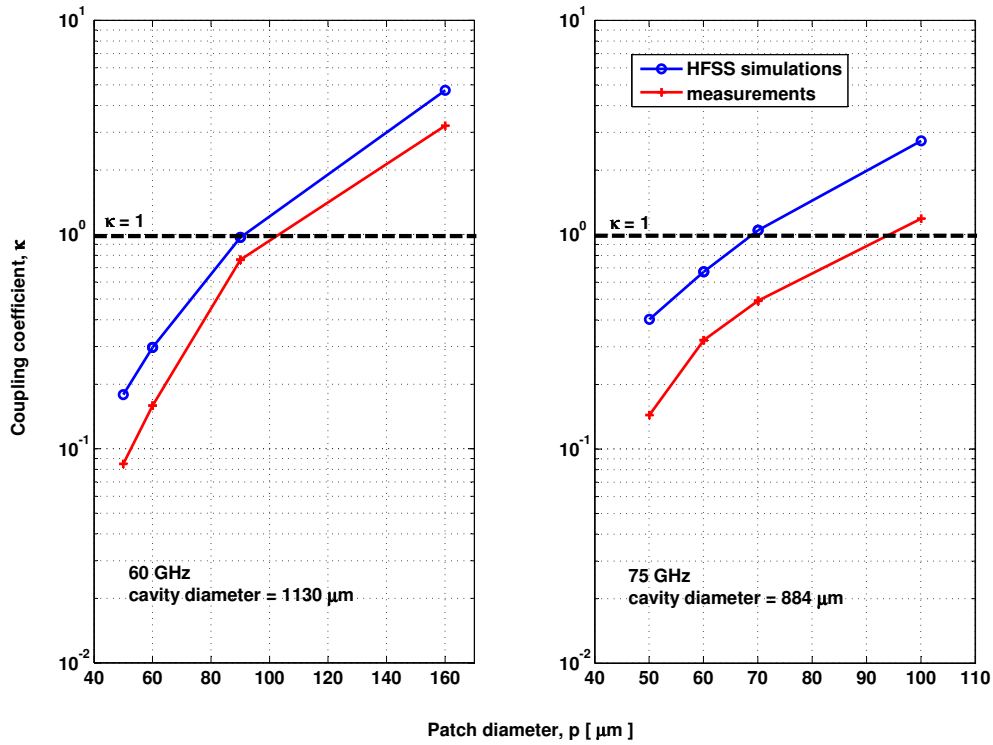


Fig. 2.32.: Simulated and measured influence of diameter of the feeding patch,  $p$  on the coupling coefficient,  $\kappa$ . The nominal diameter of the resonator is  $D = 1130 \mu m$  at  $60$  GHz and  $D = 884 \mu m$  at  $75$  GHz.

When  $\kappa = 1$ , as can be seen in Figure 2.33 for the loaded Q-factor, in Figure 2.34 for the external Q-factor and in Figure 2.35 for the unloaded Q-factor, at  $60$  GHz for the HFSS simulations:  $Q_e = 165 \approx Q_0 = 160$ ,  $Q_l = 85$  and for the measurements:  $Q_e = 135 \approx Q_0 = 130$ ,  $Q_l = 60$  at  $75$  GHz for the HFSS simulations:  $Q_e = 165 \approx Q_0 = 175$ ,  $Q_l = 80$  and for the measurements:  $Q_e = 100 \approx Q_0 = 100$ ,  $Q_l = 60$ .



Table 2.5.: At 60 and 75 GHz, simulated and measured loaded Q-factor,  $Q_l$  and coupling coefficient,  $\kappa$  as a function of the diameter of the feeding patch,  $p$ .

| $D$<br>( $\mu m$ ) | $p$<br>( $\mu m$ ) | $Q_l$ |       | $\kappa$ |       |
|--------------------|--------------------|-------|-------|----------|-------|
|                    |                    | Sim.  | Meas. | Sim.     | Meas. |
| 1130               | 50                 | 137.7 | 113.7 | 0.18     | 0.09  |
| 1130               | 60                 | 124.6 | 108.6 | 0.29     | 0.16  |
| 1130               | 90                 | 81.3  | 77.5  | 0.97     | 0.76  |
| 1130               | 160                | 28.0  | 25.8  | 4.71     | 3.22  |
| 884                | 50                 | 123.9 | 96.4  | 0.43     | 0.14  |
| 884                | 60                 | 103.8 | 79.9  | 0.67     | 0.32  |
| 884                | 70                 | 84.5  | 70.4  | 1.05     | 0.49  |
| 884                | 100                | 45.9  | 43.1  | 2.74     | 1.19  |

Table 2.6.: At 60 and 75 GHz, simulated and measured external,  $Q_e$  and unloaded,  $Q_0$  Q-factors as a function of the diameter of the feeding patch,  $p$ .

| $D$<br>( $\mu m$ ) | $p$<br>( $\mu m$ ) | $Q_e$ |        | $Q_0$ |       |
|--------------------|--------------------|-------|--------|-------|-------|
|                    |                    | Sim.  | Meas.  | Sim.  | Meas. |
| 1130               | 50                 | 908.5 | 1432.9 | 162.3 | 123.5 |
| 1130               | 60                 | 545.4 | 790.4  | 161.5 | 125.9 |
| 1130               | 90                 | 165.1 | 179.3  | 160.2 | 136.5 |
| 1130               | 160                | 34.0  | 33.8   | 159.7 | 109.1 |
| 884                | 50                 | 431.5 | 764.9  | 173.8 | 110.3 |
| 884                | 60                 | 258.4 | 328.3  | 173.5 | 105.6 |
| 884                | 70                 | 165.1 | 213.2  | 173.1 | 105.1 |
| 884                | 100                | 62.7  | 79.3   | 171.6 | 94.4  |

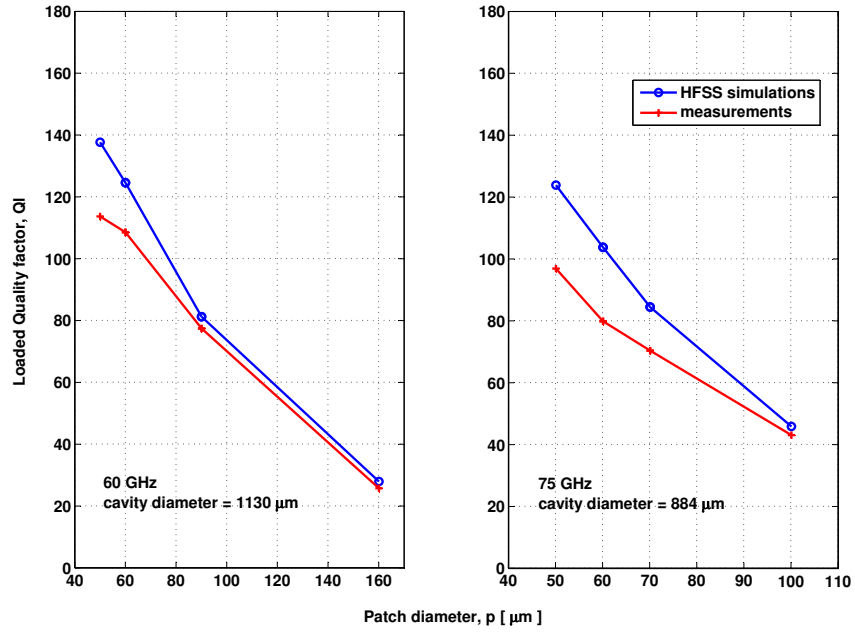


Fig. 2.33.: Simulated and measured loaded Q-factor,  $Q_l$ , as a function of  $p$ .

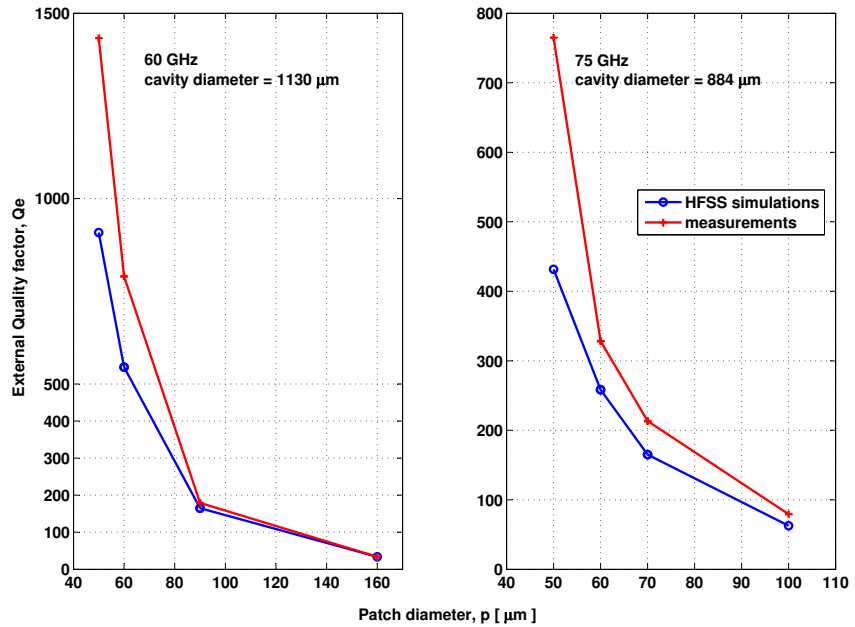


Fig. 2.34.: Simulated and measured external Q-factor,  $Q_e$ , as a function of  $p$ .

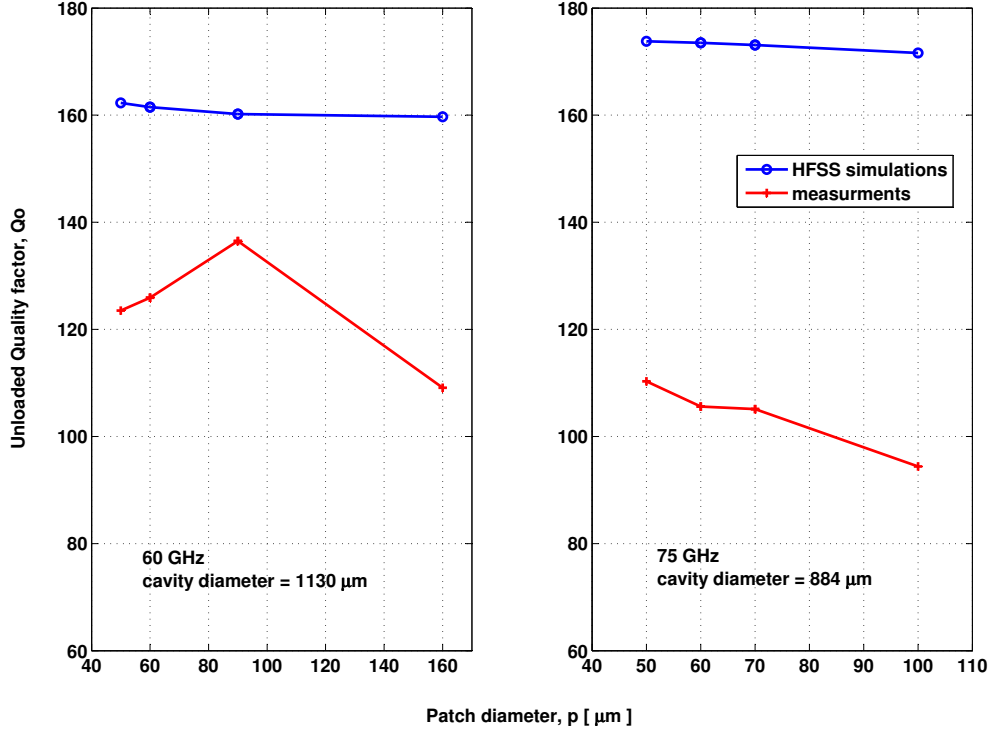


Fig. 2.35.: Simulated and measured unloaded Q-factor,  $Q_0$ , as a function of  $p$ .

It is also remarkable that the extracted Q-factors presented in Figure 2.33 and Figure 2.34 are strongly dependent on the coupling coefficient and therefore on the patch diameter while this is not the case in Figure 2.35 for the unloaded  $Q_0$ .

When  $\kappa = 0$  and  $p$  is small ( $p < 20 \mu m$ ), the resonator is not coupled and  $Q_l = Q_0$ , which represents the maximum value attainable by  $Q_l$ . For the HFSS simulations on average and at 60 GHz,  $Q_l = Q_0 \approx 160$  and at 75 GHz  $Q_l = Q_0 \approx 170$ , elsewhere  $Q_l < Q_0$  and for the measurement, on average at 60 GHz,  $Q_l = Q_0 \approx 130$  and at 75 GHz  $Q_l = Q_0 \approx 120$ , elsewhere  $Q_l < Q_0$ .

In the other undercritically coupled cases, less power is dissipated in the external circuit than in the resonator and therefore  $Q_e > Q_0$  and  $Q_0/2 < Q_l < Q_0$ .

While for an overcritical coupling, more power is lost in the external circuit than in the resonator and  $Q_e < Q_0$  and therefore  $Q_l < Q_0/2 < Q_0$ .

## 2.6 EXTRACTION OF THE UNLOADED RESONANCE FREQUENCY

A new empirical method allows the extraction of the unloaded resonance frequency,  $f_0$  from a set of at least three measurements [Canos 03]-[Canos 06]. The extraction is realized without the knowledge of the electromagnetic field distribution neither in the

resonator nor with the knowledge of the exact group delay of the feeding transmission lines and coupling mechanism. When  $Q_e$  takes very high values, the loaded resonance frequency,  $f_r$  tends towards  $f_0$  and the relation between  $f_r$  and  $f_0$  is given as follows [Canos 03]-[Canos 06]:

$$f_r = \left(1 + \frac{A}{Q_e^\alpha}\right) f_0 \quad (2.92)$$

with the unknown parameters  $A$ ,  $\alpha$  and  $f_0$ . These parameters are empirically determined and are dependent on the electromagnetic behavior of the loaded resonator, feeding included. The parameters are computed from a system of at least three equations, accordingly to [Canos 03]-[Canos 06]:

$$f_{r,i} = \left(1 + \frac{A}{Q_{e,i}^\alpha}\right) f_0, \quad i = 1, 2, \dots, N; \quad N \geq 3 \quad (2.93)$$

The system of equations is composed indifferently of simulations or measurements, but the resonators should be differentiated by the feeding structure and therefore characterized by different external Q-factors,  $Q_e$ . Better results are obtained with a family of resonators sweeping coupling coefficients in an homogenous increasing way, from undercoupled coupling towards overcoupled coupling, passing through a critically coupled resonator.

In our case, four different systems, each composed by four equations are individually solved to extract the unloaded resonance frequency,  $f_0$ , one per system. The four elements in each system involve different  $Q_e$ , as each resonator is having a different patch diameter, hence different  $f_r$ . Simulations and measurements are considered at 60 GHz ( $D = 1130 \mu m$ ) and at 75 GHz ( $D = 884 \mu m$ ), yielding the four systems. The geometrical parameters of the resonators are given in Table 2.5 and in Table 2.6. The simulations and measurements are used to extract the unknown parameters  $A$ ,  $\alpha$  and  $f_0$  of (2.93) and are presented in Table 2.7.

Table 2.7.: Extracted parameters  $A$ ,  $\alpha$  and  $f_0$  from both simulations and measurements at 60 and 75 GHz.

| $D (\mu m)$ |       | $A$    | $\alpha$ | $f_0$ (GHz) |
|-------------|-------|--------|----------|-------------|
| 1130        | Sim.  | 1.4564 | 1.3677   | 58.9551     |
| 1130        | Meas. | 0.6769 | 1.1793   | 59.2613     |
| 884         | Sim.  | 1.2510 | 1.3822   | 75.1779     |
| 884         | Meas. | 0.4761 | 1.0491   | 74.7067     |

Those parameters are used to calculate the detuning of the loaded resonance frequency,  $f_r$  due to the coupling, presented in Figure 2.36, using (2.92). In this figure,  $f_0$  is

the asymptotic limit towards which the loaded resonance frequency,  $f_r$ , tends as the external Q-factor,  $Q_e$  grows.

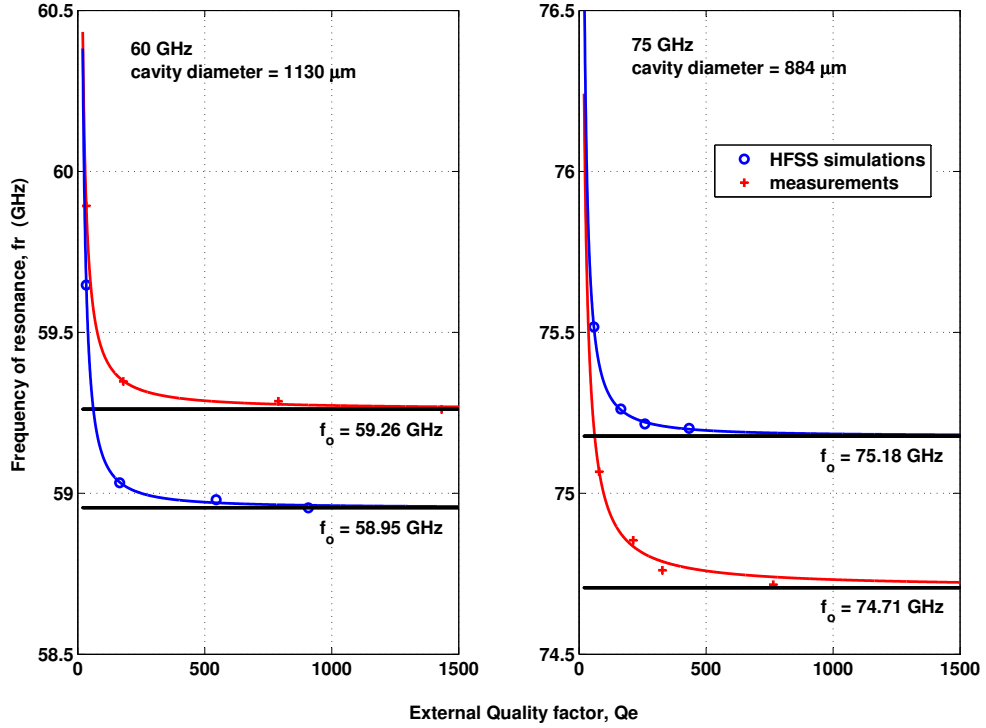


Fig. 2.36.: Simulated and measured influence of the external Q-factor,  $Q_e$  on the loaded resonance frequency,  $f_r$ . The diameter of the resonator is  $D = 1130 \mu\text{m}$  at 60 GHz and  $D = 884 \mu\text{m}$  at 75 GHz.

## 2.7 EQUIVALENT CIRCUIT EXTRACTION

According to Foster's formalism, presented in [Kajfez 94], the detuning due to the coupling of a cavity resonator is described by a series impedance  $jX_s$  with a parallel RLC tank, in Figure 2.37-a) or by a shunt admittance  $jB_e$  with a series RLC tank, in Figure 2.37-b). The Ohmic losses are neglected, while  $X_s$  and  $B_e$  represent the extra energy stored in the coupling structure, accounting for the detuning of the measured resonance frequency. The transformer coefficient,  $n$  realizes the modelling of the coupling and assumes impedance adjustment.

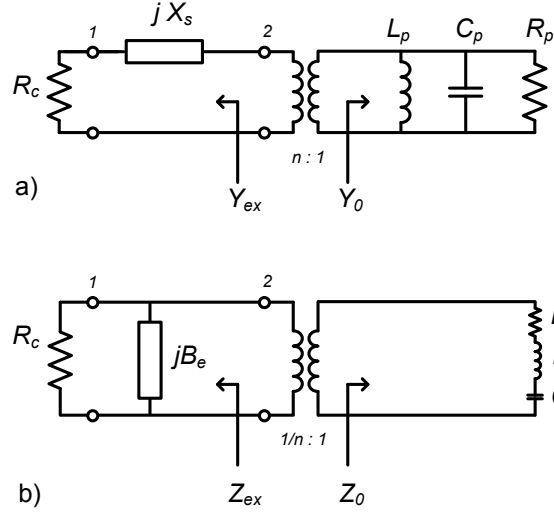


Fig. 2.37.: a) First Foster's form, parallel RLC lumped model of the loaded resonant cavity, with series reactance  $X_s$ . b) Second Foster's form, series RLC lumped model of the loaded resonant cavity, with parallel susceptance  $B_e$ .

The choice of the correct Foster's form is based on both the sign of the reflection coefficient out of resonance,  $\Gamma_d$  and the sign of the difference between the loaded and unloaded resonance frequency,  $(f_r - f_0)$ , following [Canos 06] in Table 2.8.

Table 2.8.: Criterion for Foster's form choice

| $Im\{\Gamma_d\}$ | $(f_r - f_0)$ | Equivalent resonant Circuit |           |
|------------------|---------------|-----------------------------|-----------|
| $> 0$            | $> 0$         | Parallel RLC                | $X_s > 0$ |
| $> 0$            | $< 0$         | Serial RLC                  | $B_e < 0$ |
| $< 0$            | $> 0$         | Serial RLC                  | $B_e > 0$ |
| $< 0$            | $< 0$         | Parallel RLC                | $X_s < 0$ |

In the first Foster's form, the normalized out of resonance reflection coefficient,  $\Gamma_d$  is given by (2.94) where  $X_s$  is normalized to the characteristic impedance  $Z_c = 50 \Omega$ :

$$\Gamma_d = \frac{jX_s - 1}{jX_s + 1} \quad (2.94)$$

while in the second Foster's model,  $\Gamma_d$ , the detuned reflection coefficient is given by (2.95), where  $B_e$  is normalized to the characteristic impedance  $Z_c = 50 \Omega$ :

$$\Gamma_d = \frac{1 - jB_e}{1 + jB_e} \quad (2.95)$$

The detuning of the loaded resonance frequency,  $f_r$ , is given in the first Foster's model, as follows:

$$f_r = \left(1 + \frac{jX_s}{2Q_e}\right) f_0 \quad (2.96)$$

while in the second Foster's model,  $f_r$  is given as:

$$f_r = \left(1 + \frac{jB_e}{2Q_e}\right) f_0 \quad (2.97)$$

The unfolding of the criterion presented in Table 2.8 for the choice of the appropriate Foster's Form, comes as follows:

When  $Im\{\Gamma_d\} < 0$ , either  $X_s < 0$  in (2.94) then  $(f_r - f_0) < 0$ , following (2.96) or  $B_e > 0$  in (2.95) then  $(f_r - f_0) > 0$ , following (2.97).

Reciprocally, when  $Im\{\Gamma_d\} < 0$  and  $(f_r - f_0) < 0$  the first Foster's form is used with  $X_s < 0$  while when  $Im\{\Gamma_d\} < 0$  and  $(f_r - f_0) > 0$ , the second Foster's form is used with  $B_e > 0$ .

When  $Im\{\Gamma_d\} < 0$ , either  $B_e < 0$  in (2.95) then  $(f_r - f_0) < 0$ , following (2.96) or  $X_s > 0$  in (2.94) then  $(f_r - f_0) > 0$ , following (2.97).

Reciprocally, when  $Im\{\Gamma_d\} > 0$  and  $(f_r - f_0) < 0$  the second Foster's form is used and  $B_e < 0$  while when  $Im\{\Gamma_d\} > 0$  and  $(f_r - f_0) > 0$ , the first Foster's form is used and  $X_s > 0$ .

The criterion for the choice of the Foster form is presented in Table 2.8. It depends on the sign of both  $Im\{\Gamma_d\}$  and  $(f_r - f_0)$ . The transition from the first Foster's form to the second one is realized considering the quarter wavelength transformation,  $Z = R_c^2 Y_p$  applied at port 1 in Figure 2.38-a). This transformation is equivalent to re-distributing quarter wavelength transformations to each of the model components, till the parallel RLC which is directly transformed into the serial RLC, as described in Figure 2.38-b). As a consequence of the redistribution, the intermediate equivalent components of the second Foster's model are calculated following circuits theory, using ABCD matrixes applied at each intermediated bloc. The detail of the calculation is presented in Appendix A and the final results are given as follows:

$$R_s = R_c^2 / R_p \quad (2.98)$$

$$C_s = L_p / R_c^2 \quad (2.99)$$

$$L_s = R_c^2 C_p \quad (2.100)$$

$$B_e = X_s / R_c^2 \quad (2.101)$$

$$n_s = \frac{1}{n_p} \quad (2.102)$$

with the characteristic impedance  $R_c = 50 \Omega$ .

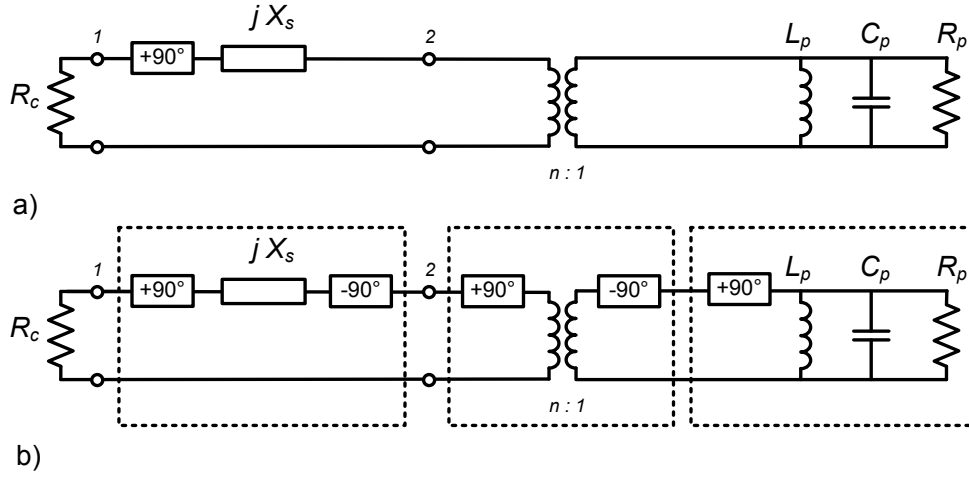


Fig. 2.38.: a) Transition from the first Foster's form to the second form. b) Distributed transition from the first Foster's form to the second form, details of the calculation are presented in Appendix A.

The transformer coefficient,  $n$  of the first Foster's form renamed,  $n_p = n$  is inverted in second Foster's form,  $n_s = 1/n$  and is calculated using the coupling coefficient,  $\kappa$  extracted in Section 2.5.3, as follows:

$$n = \sqrt{\frac{\kappa}{\frac{R_c}{R_c^2 + X_s^2} R_p}} \quad (2.103)$$

with  $R_p$  the shunt impedance in first Foster's form. The equivalent shunt impedance,  $R_p$  is calculated using the form factor,  $FF_{TM010}$ , as follows:

$$R_p = Q_0 FF_{TM010} \quad (2.104)$$

The form factor is only function of the physical dimensions of the resonator, dependency which can be rewritten as a function of the resonance frequency and height of the resonator, adapted from [Schmuser 03], as follows:

$$FF_{TM010} = \frac{8H_{HR-Si}\mu_0\sqrt{\epsilon_r}}{(\pi J_1(2.405)2.405)^2} f_0 \quad (2.105)$$



By using (2.104), all sources of losses are taken into consideration as the extracted unloaded Q-factor,  $Q_0$  is defined as (2.23).

In the first Foster's form the loaded resonance frequency is given in [Kajfez 94]:

$$f_r = \left(1 + \frac{\kappa X_s}{2Q_0 R_c}\right) \quad (2.106)$$

The admittance  $X_s$  is also function of the parameters  $A$  and  $\alpha$  considering (2.92) and (2.106), as follows:

$$X_s = \frac{A}{Q_e^\alpha} \frac{2Q_0 R_c}{\kappa} \quad (2.107)$$

with  $Q_0$ ,  $Q_e$  and  $\kappa$  extracted in Section 2.5.3

The elements  $C_p$  and  $L_p$  are calculated as follows:

$$C_p = \frac{1}{2\pi f_0} \frac{1}{FF_{TM010}} \quad (2.108)$$

$$L_p = \frac{1}{2\pi f_0} FF_{TM010} \quad (2.109)$$

As a consequence, the parallel inductance,  $L_p$  or similarly the serial capacitance,  $C_s$  are independent of the resonance frequency, considering (2.109) and (2.105). They depend finally only on the height of the cylindrical resonator,  $H_{HR-Si}$  and permittivity of the filling material (2.90), for the given mode of resonance, i.e.  $TM_{010}$ .

As both the measurements, as well as the HFSS simulations are characterized by  $Im\{\Gamma_d\} < 0$  and  $(f_r - f_0) > 0$  (Figure 2.36 and Figure 2.41-2.44), the second Foster's form is used to describe the detuning of the loaded cavity resonator with the equivalent parameters calculated following the equations (2.98)-(2.101), following the quarter wave length transformation, from the first to the second form, described in Figure 2.38.

The parameters of the second Foster's form, extracted from measurements are presented in the Table 2.9 and those extracted from HFSS simulations in Table 2.10. They are provided in two families of resonators, each family being composed of four different resonators characterized by different diameters of the feeding patch,  $p$ , corresponding respectively to resonators at 60 and 75 GHz. In these tables are given the loaded resonance frequency, the transformer coefficient, the susceptance and the series resistance.

The series capacitance is independent of the resonance frequency, but following (2.99) and (2.109) only function of the height of the resonator. Using those equations, the extracted series capacitance is  $C_s = 15.282 \text{ fF}$  for measurements and simulations at both 60 and 75 GHz. Using (2.100) and (2.108), for the measurements at 60 GHz, the inductance,  $L_s = 0.472 \text{ nH}$ , therefore the unloaded resonance frequency,  $f_0 = 59.2613 \text{ GHz}$  and on average,  $R_s = 1.429 \Omega$  following (2.98) and (2.104). Using the same pair of equations, for the measurements at 75 GHz,  $L_s = 0.297 \text{ nH}$ ,

$f_0 = 74.7067 \text{ GHz}$  and on average,  $R_s = 1.347 \Omega$ . For the simulations at 60 GHz,  $L_s = 0.477 \text{ nH}$ ,  $f_0 = 58.9551 \text{ GHz}$  and on average,  $R_s = 1.098 \Omega$ . For the simulations at 75 GHz,  $L_s = 0.293 \text{ nH}$ ,  $f_0 = 75.1779 \text{ GHz}$  and on average,  $R_s = 0.798 \Omega$ .

Table 2.9.: Parameters of the lumped model extracted from measurements.

| $D$<br>( $\mu m$ ) | $p$<br>( $\mu m$ ) | $f_r$<br>(GHz) | $n$    | $B_e$<br>( $mS$ ) | $R_s$<br>( $\Omega$ ) |
|--------------------|--------------------|----------------|--------|-------------------|-----------------------|
| 1130               | 50                 | 59.26          | 0.0545 | 7.42              | 1.530                 |
| 1130               | 60                 | 59.29          | 0.0747 | 8.19              | 1.500                 |
| 1130               | 90                 | 59.35          | 0.1646 | 10.66             | 1.384                 |
| 1130               | 160                | 59.89          | 0.4120 | 14.42             | 1.732                 |
| 884                | 50                 | 74.75          | 0.0760 | 13.75             | 1.359                 |
| 884                | 60                 | 74.76          | 0.1176 | 14.30             | 1.419                 |
| 884                | 70                 | 74.85          | 0.1469 | 14.66             | 1.426                 |
| 884                | 100                | 75.07          | 0.2451 | 15.37             | 1.587                 |

Table 2.10.: Parameters of the lumped model extracted from HFSS simulations.

| $D$<br>( $\mu m$ ) | $p$<br>( $\mu m$ ) | $f_r$<br>(GHz) | $n$    | $B_e$<br>( $mS$ ) | $R_s$<br>( $\Omega$ ) |
|--------------------|--------------------|----------------|--------|-------------------|-----------------------|
| 1130               | 50                 | 58.96          | 0.0665 | 4.76              | 1.170                 |
| 1130               | 60                 | 58.98          | 0.0869 | 5.73              | 1.176                 |
| 1130               | 90                 | 59.03          | 0.1660 | 8.92              | 1.185                 |
| 1130               | 160                | 59.65          | 0.4277 | 15.91             | 1.189                 |
| 884                | 50                 | 75.22          | 0.0866 | 6.66              | 0.792                 |
| 884                | 60                 | 75.24          | 0.1130 | 8.05              | 0.793                 |
| 884                | 70                 | 75.29          | 0.1425 | 9.51              | 0.796                 |
| 884                | 100                | 75.49          | 0.2364 | 13.91             | 0.803                 |

The smaller the diameter of the feeding patch,  $p$ , the lower the coupling coefficient,  $\kappa$ . As a consequence, the transformer coefficient,  $n$  decreases, as can be seen in Figure 2.39 and is approximated by a second order polynomial with intercept in (0,0), provided as a function of  $p$ :

$$n(p) = \alpha_1 p^2 + \alpha_2 p \quad (2.110)$$

The dependency of  $n$  to the square of the diameter is related to the surface of the coupling, following Wheeler's concept of effective volume ratios for evaluating the couplings between cavity resonators [Wheeler 63].

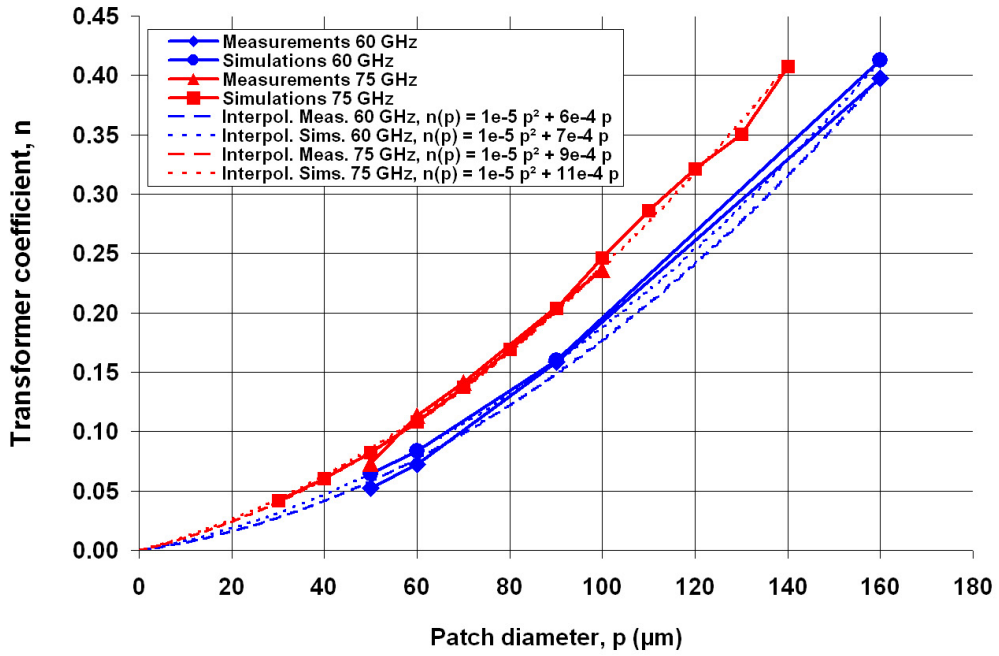


Fig. 2.39.: Transformer coefficient,  $n$  extracted from measurements and HFSS simulations, as a function of the patch diameter,  $p$ .

Simultaneously, the phase shift induced by the coupling, modelled by the shunt susceptance,  $B_e$  also decreases, as can be seen in Figure 2.40 and the susceptance is approximated linearly with intercept in (0,0), as a function of  $p$ :

$$B_e(p) = \alpha_3 p \quad (2.111)$$

The parameters  $\alpha_1$ ,  $\alpha_2$  and  $\alpha_3$  are empirically extracted from measurements and simulations at both 60 and 75 GHz and are presented in Table 2.11. They allow writing a direct approximation of the transformer coefficient and shunt susceptance as a function of the feeding patch diameter. The coefficients in Table 2.11 correspond to the patch diameter in ( $\mu m$ ).

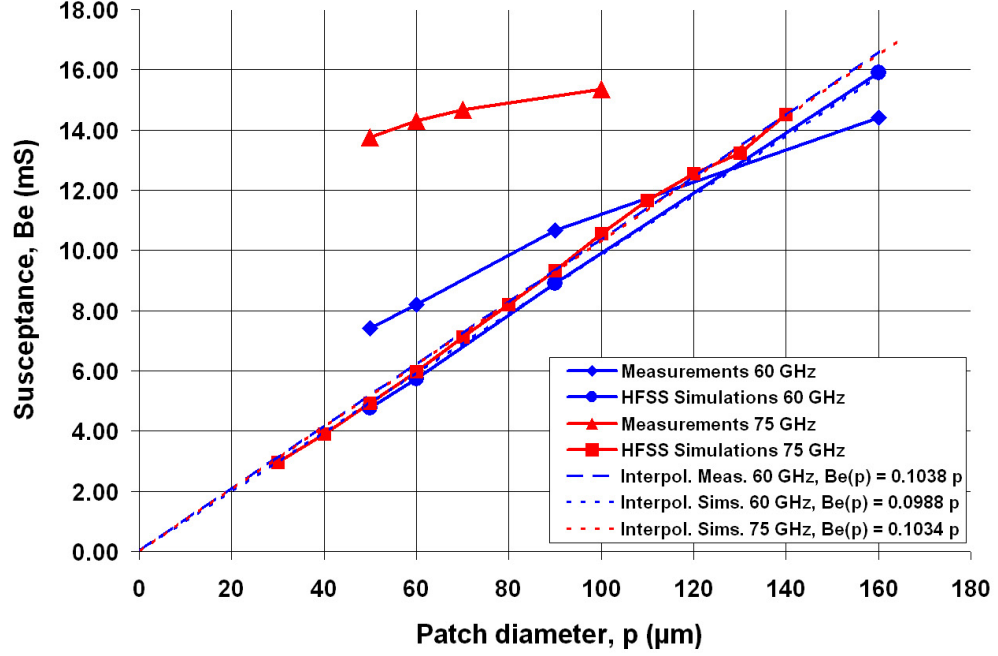


Fig. 2.40.: Susceptance,  $B_e$  extracted from measurements and HFSS simulations, as a function of the patch diameter,  $p$ .

Table 2.11.: Extracted parameters  $\alpha_1$ ,  $\alpha_2$  and  $\alpha_3$  approximating the transformer coefficient  $n$  and susceptance  $B_e$ .

|            | Meas.<br>60 GHz | Sim.<br>60 GHz | Meas.<br>75 GHz | Sim.<br>75 GHz |
|------------|-----------------|----------------|-----------------|----------------|
| $\alpha_1$ | 1e-5            | 1e-5           | 1e-5            | 1e-5           |
| $\alpha_2$ | 6e-4            | 7e-4           | 9e-4            | 11e-4          |
| $\alpha_3$ | 0.1038          | 0.0988         | -               | 0.1034         |

The comparison between the equivalent circuit and the measurements performed at 60 GHz and at 75 GHz are shown respectively in Figure 2.41 and in Figure 2.42. The measurements are de-embedded by  $16^\circ$  at 60 GHz and by  $24^\circ$  at 75 GHz, which corresponds to the phase shift of the connecting micro-strip line, with a length of  $330 \mu\text{m}$ , which is additional to the RLC model and coupling circuit  $(X_s, B_e)$ .

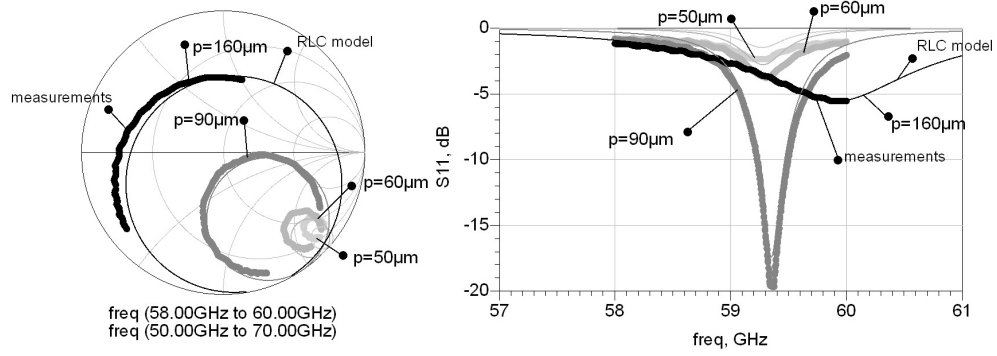


Fig. 2.41.: Comparison of the RLC model with the measurements at 60 GHz. A sampling set of four resonators is shown, with the following values for the patch diameter,  $p$ : 50, 60, 90 and 160  $\mu m$ .

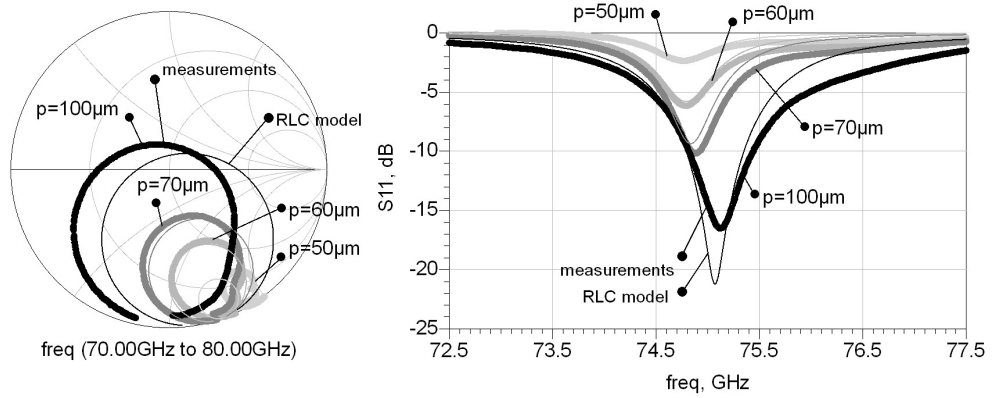


Fig. 2.42.: Comparison of the RLC model with the measurements at 75 GHz. A sampling set of four resonators is shown, with the following values for the patch diameter,  $p$ : 50, 60, 70 and 100  $\mu m$ .

A large discrepancy is visible as a phase shift for the model of the 75 GHz measured resonators, presented in Figure 2.42. This is redundant in the susceptance  $B_e$  behavior, presented in Figure 2.40, which is not pointing towards (0,0). This discrepancy is due to the extraction of the unloaded resonance frequency, parameter  $f_0$ , from the fitting system (2.93) affecting the extraction of both the reactance,  $X_s$  (2.96) and susceptance,  $B_e$  (2.97) of respectively first and second Foster's form. If the fitting is improved, the susceptance is more accurately evaluated. Actually, to improve the fitting, higher values of external Q-factors,  $Q_e$  are required. These values correspond to very small diameters of the patch feed,  $p$ . Small manufacture errors could deviate from the ideal behavior of the resonator. Measurement errors could be generated by the close proximity of the RF probes, being comparable in dimensions with the feeding patch of the cavity resonator.

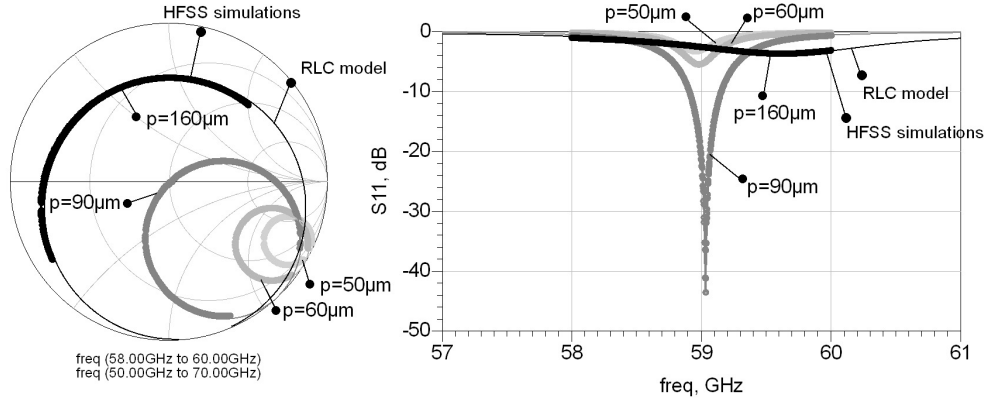


Fig. 2.43.: Comparison of the RLC model with the simulations at 60 GHz. A sampling set of four resonators is shown, with the following values for the patch diameter,  $p$ : 50, 60, 90 and 160  $\mu m$ .

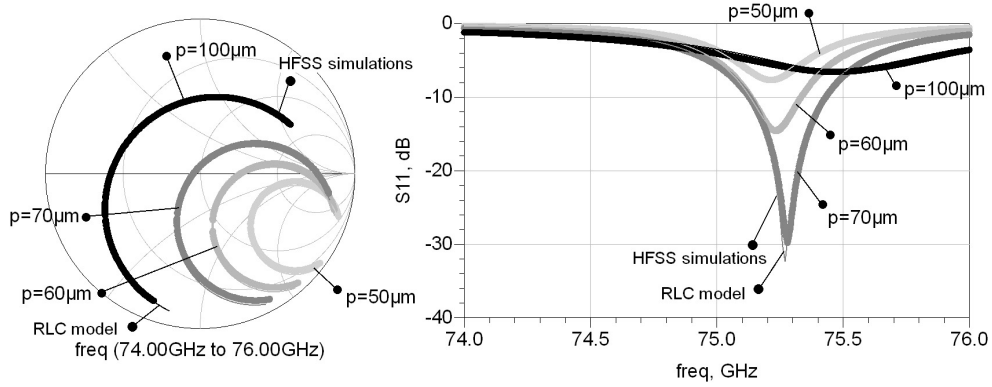


Fig. 2.44.: Comparison of the RLC model with the simulations at 75 GHz. A sampling set of four resonators is shown, with the following values for the patch diameter,  $p$ : 50, 60, 70 and 100  $\mu m$ .

The comparison between the equivalent circuit and the simulations with HFSS, performed at 60 GHz and at 75 GHz, are given respectively in Figure 2.43 and in Figure 2.44. The simulations are de-embedded by  $24^\circ$  at 60 GHz and by  $32^\circ$  at 75 GHz, which corresponds to the phase shift of the connecting micro-strip line, with a length of 330  $\mu m$ , which is additional to the RLC model and coupling circuit ( $X_s, B_e$ ). In general, the fitting by the RLC models are very well modelling the behavior of the cavity resonators simulated with HFSS, as these last ones do not present the non-idealities observed for the measurements.

## 2.8 HIGH IMPEDANCE SURFACE: ROUTE FOR CAVITY RESONATORS MINIATURIZATION

The investigation on cavity resonators miniaturization represents a way of reducing costs by reducing weight, interesting for example, in the case of placement on orbit for satellites' payload. Actually, the satellites still carry metallic waveguides based devices such as filters, diplexers and interconnections, because lighter alternatives are underperforming [Coirault 92]. Recently, theoretical developments proposed metamaterials to reduce the size of cavity resonators [Engheta 02] and waveguides [Hrabar 05]. Another approach demonstrated TEM propagation in a waveguide composed of Artificial Magnetic Conductors (AMC) [Yang 99]. Similarly, are proposed High Impedance Surfaces (HIS), electromagnetic structures showing a high impedance and in-phase reflection [Sievenpiper 99]. HIS are used to reduce antennas profiles [Mateos 09], improve the radiation pattern by suppressing surface waves [Sievenpiper 99] and for beam steering [Chicherina 08]. In this context, the concept we demonstrate in this section consists in shaping a quarter wavelength long cavity resonator by using a combination of Perfect Electric Conductor (PEC) and Perfect Magnetic Conductor (PMC) boundary conditions.

The original idea, presented here, is to create a quarter wavelength long cavity resonator by using a combination of Perfect Electric Conductor (PEC) and Perfect Magnetic Conductor (PMC) boundary conditions. The length and width of a metallic cavity resonator are reduced by a factor two by placing PMC boundary conditions on two adjacent side walls. This configuration allows the resonance of a quarter wavelength mode inside the hybrid cavity resonator, at the same frequency as the idle metallic counterpart. The practical implementation of the PMC boundary condition is realized by using High Impedance Surfaces (HIS).

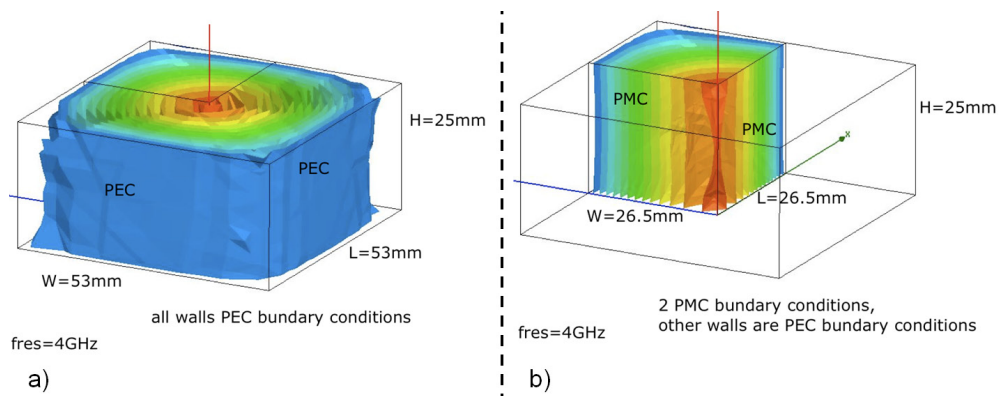


Fig. 2.45.: First resonant mode  $TE_{101}$  at 4 GHz in a) full PEC cavity and in b) hybrid PEC-PMC cavity, with respective dimensions:  $a \times d \times b = 53 \times 53 \times 25$  mm and  $26.5 \times 26.5 \times 25$  mm.

This hybrid configuration allows to reduce by a factor two the length and width of a metallic cavity resonator by placing PMC boundary conditions on two adjacent side walls. This allows the resonance of a quarter wavelength mode inside the hybrid cavity resonator, at the same frequency as the half wavelength long metallic counterpart, idle cavity resonator. More specifically, a metallic cavity resonator is characterized by a field distribution related to the modes of resonance and geometry of the resonator. The boundary conditions at all ends are assumed to be PEC and impose a zero of the electric field. Considering a rectangular metallic cavity of dimensions  $d$ ,  $a$ , and  $b$ , along the x, y, and z-axes, see Figure 2.45-a), filled with a material of permittivity  $\epsilon_r$  and permeability  $\mu_r$ , the resonance frequency is given by [Pozar 05] and referred in Section 2.4.

Instead of using PEC boundary conditions spaced by half a waveguide length, a combination of PEC and PMC boundary conditions spaced by a quarter waveguide length are used to maintain the same field distribution e.g. *maxima* and *minima* as can be seen in Figure 2.45-b). In this case the resonance frequency is given by:

$$f_{mnl} = \frac{k_{mnl}}{2\pi\sqrt{\mu\epsilon}} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{(m-1/2)\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{(l-1/2)\pi}{d}\right)^2} \quad (2.112)$$

with the fundamental mode TE<sub>101</sub> as lowest resonance frequency:

$$f_{101} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{\pi}{2a}\right)^2 + \left(\frac{\pi}{2d}\right)^2} \quad (2.113)$$

Comparison of (2.112) and (2.113) shows that the cavity with two PMC sidewalls resonates at half the frequency of a cavity having the same dimensions and all PEC sidewalls. Equivalently, since the TE<sub>101</sub> field distribution corresponds to a quarter-wavelength instead of a half-wavelength, the volume of the hybrid PEC-PMC design is a quarter of the volume of a standard PEC rectangular design, for a specified resonance frequency, as can be seen Figure 2.45-b).

The practical implementation of the PMC boundary condition is realized using High Impedance Surfaces (HIS). The design of the surfaces is realized at the element level and is implemented on standard microwave substrate. Equivalently keeping all dimensions constant, the cavity with HIS sidewalls resonates at half the frequency of a full metallic one. We measured the resonance of a fabricated hybrid HIS-metallic cavity resonator to be 4 GHz while measured without the HIS it was found to be 8 GHz.

### 2.8.1 High Impedance Surface design

High Impedance Surfaces (HIS) are known to show, around the design frequency, a PMC behavior. While the metallic sheet imposes a phase reversal (180°) for reflected plane waves, HIS imposes no (0°) phase shift while reflecting plane waves. This surface is realized by gathering together metallic patches elements visible in Figure 2.46.



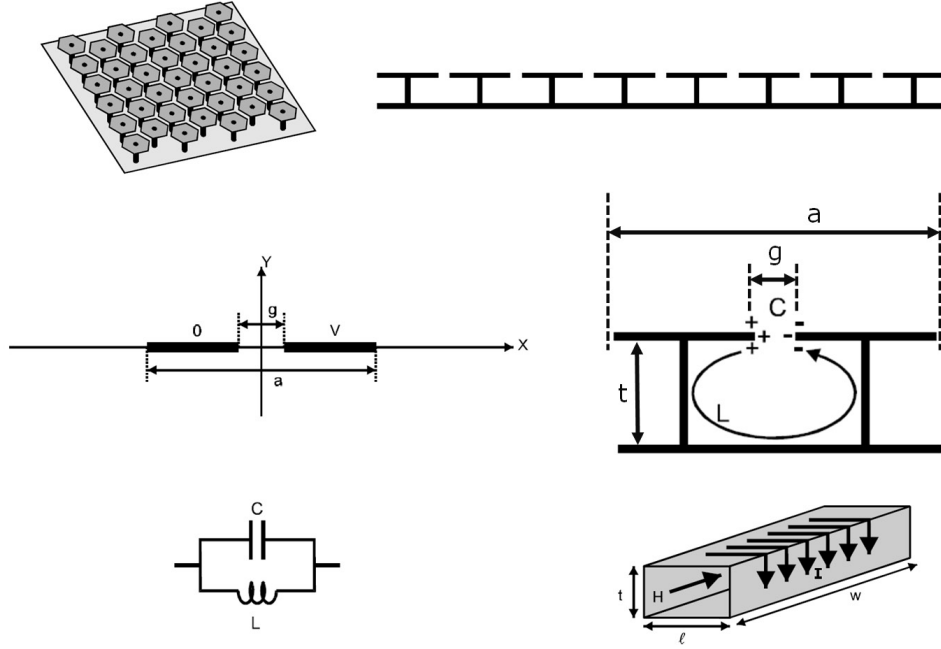


Fig. 2.46.: Mushroom HIS structure and equivalent circuit [Sievenpiper 99].

Characteristic for AMC surfaces is the zero surface impedance, at resonance, determined by the ratio of in plane electric field to magnetic field. The behavior of each element is described using lumped parameter circuit model shown as inset in Figure 2.46. The value of each capacitor is given by the fringing field capacitance between neighboring metal plates, derived using conformal mapping [Sievenpiper 99]. The inductance is strongly dependent of the substrate thickness, forming a solenoid of current defined on one side by the metallic patch and on the other side by the walls of the cavity in which the PCB is included. The magnetic energy stored in the inductor is directly proportional to the volume of the solenoid wherein the magnetic field engulfs, generating surface currents flowing around. Values for the capacitance and inductance of a single patch element are presented in [Sievenpiper 99], with  $\varepsilon_1$  and  $\varepsilon_2$  the permittivity of dielectric and air,  $\mu$  the permeability and the thickness of the PCB substrate,  $t$ :

$$C = \frac{W(\varepsilon_1 + \varepsilon_2)}{\pi} \cosh^{-1} \left( \frac{\text{slot}}{2W + \text{slot}} \right) \quad (2.114)$$

$$L = \mu t \quad (2.115)$$

In our implementation, the patches are lying on the microwave substrate and are enclosed by the metallic walls of the cavity. The HIS element is enclosed in PEC

boundary conditions at the rear, the top and bottom. On both lateral sides PMC boundary conditions are placed at half the slot length, in order to render the periodicity of the HIS. The element is excited by a plane wave with vertical  $E_z$  field and horizontal  $H_x$  field. The resonance frequency is obtained from HFSS simulations at the frequency where the phase of reflection coefficient vanishes, as it is shown in Figure 2.47-a). The patch element can further be schematically represented in Figure 2.47-b,-c), where a disconnected patch is enclosed by the metallic walls of the cavity resonator.

Further finite elements HFSS simulations were performed using the configurations presented in Figure 2.47.

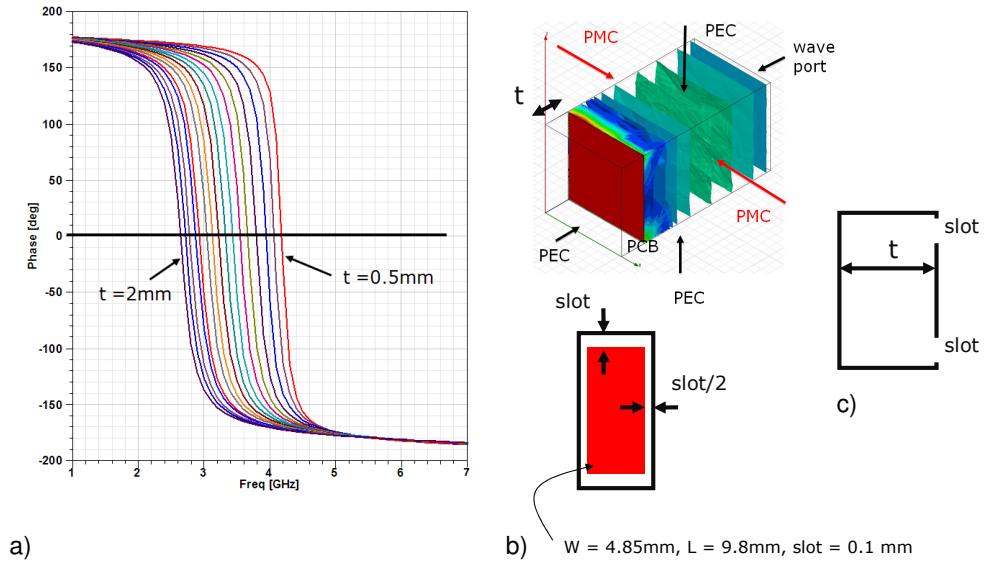


Fig. 2.47.: a) Phase of reflection coefficient  $S_{11}$  simulated using HFSS software for a unit cell composing the high impedance surface. b) E field distribution in the simulation setup. The phase is computed versus frequency for various thicknesses  $t$  of the substrate, varying from 0.5 to 2 mm. c) Cross-section of the HIS element showing the thickness of the PCB and the slots on top and bottom of the patch disconnecting it from the inner cavity's faces.

Two prototypes, shown in Figure 2.48 were investigated. In the case of the first prototype, (2.114-2.115) are adapted to the characteristic geometry presented in Figure 2.48-b)-c). In this configuration, the HIS element is composed by the cavity's metallic walls, the substrate and a metallic patch. The patch is spaced by a slot from the top and bottom of the cavity's walls. A lumped circuit is derived (see inset Figure 2.48-b)) wherein two series equivalent inductors and two series equivalent capacitors are identified. The resonance frequency of each HIS element of the first kind is given by the following equation:

$$f_{0,HIS1} = \frac{1}{2\pi\sqrt{2L_2^C}} \quad (2.116)$$

The characteristic dimensions of each metallic patch are  $W_{HIS1} = 4.85$  mm,  $L_{HIS1} = 9.8$  mm, spaced by a slot of  $100 \mu m$  from the top and bottom of the cavity walls. Also in Figure 2.48-d) is presented the comparison between the resonance frequency as calculated using (2.116) and resulting from HFSS simulations. Both methods are in good agreement and suggests a thickness,  $t_{HIS1} = 0.7$  mm in order to establish the resonance of each element at 4 GHz.

In the case of the second prototype, (2.114-2.115) are adapted to the characteristic geometry presented in Figure 2.48-e)-g), where the HIS element is composed by the cavity's walls, the substrate and two metallic patches. The patches are spaced by a slot from the top and bottom of the cavity's walls and there is a distance of a slot in between the two patches. A lumped circuit is derived (see inset Figure 2.48-f)) wherein two series equivalent inductors and three series equivalent capacitors are identified. The resonance frequency of each HIS element of the second kind is given by the following equation:

$$f_{0,HIS2} = \frac{1}{2\pi\sqrt{2L_3^C}} \quad (2.117)$$

The characteristic dimensions of each metallic patch are  $W_{HIS2} = 4.85$  mm,  $L_{HIS2} = 4.85$  mm, slot =  $100 \mu m$ . Two patches are spaced by a slot of  $100 \mu m$  and also spaced by the same amount from the top and bottom of the cavity walls. Also in Figure 2.48-h) is the comparison between the resonance frequency as calculated using (2.117) and resulting from HFSS simulations. Both methods are in good agreement and suggests a thickness,  $t_{HIS2} = 2.5$  mm in order to establish the resonance of each element at 4 GHz.

A good agreement is observed between the resonant frequencies calculated for the configuration proposed using respectively, the lumped element approach of formulas (2.114-2.115) (solid black lines) and the full wave simulations with HFSS (dotted red lines) as can be seen in Figure 2.48-h).

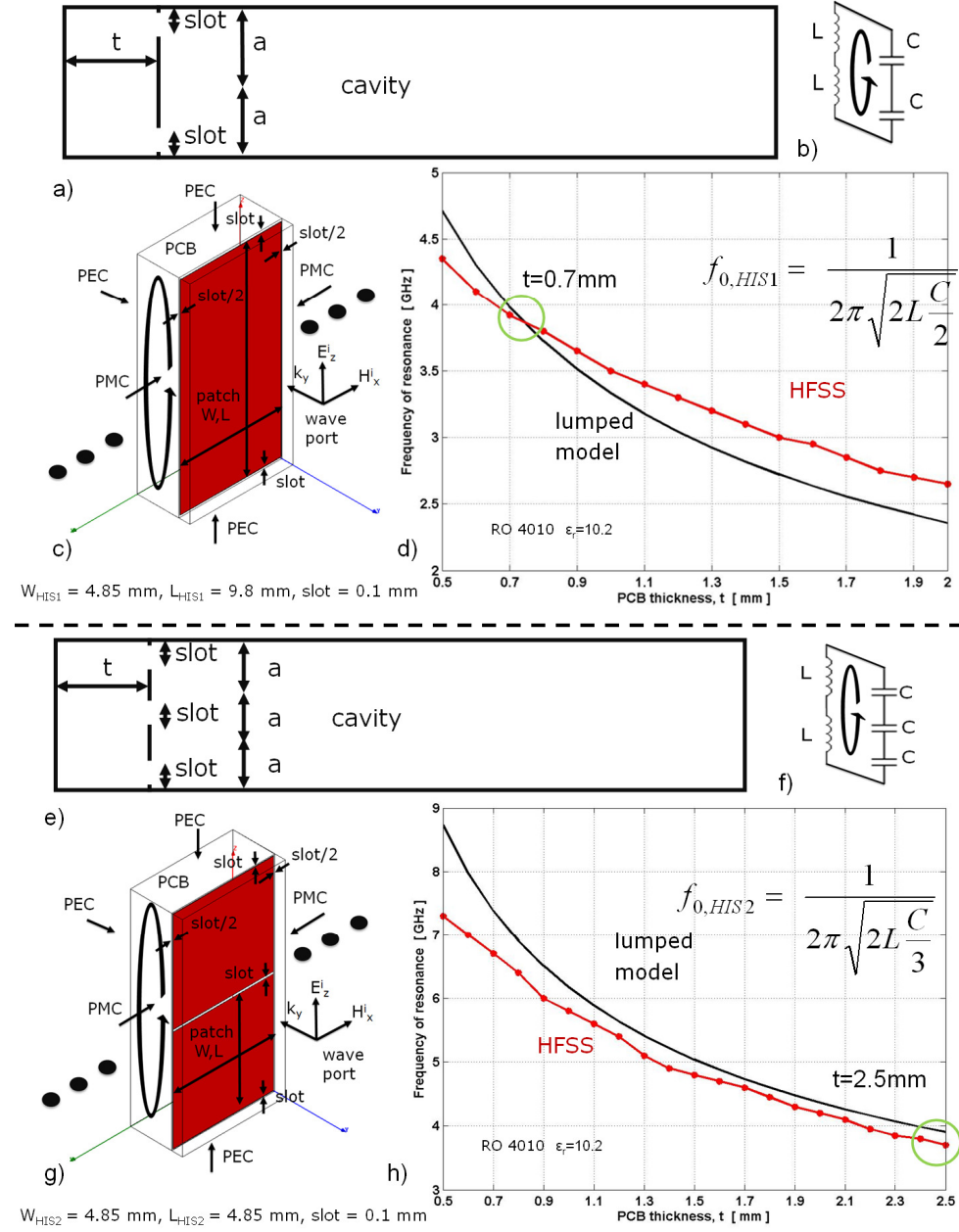


Fig. 2.48.: First and second HIS under study: a,e) each element of the HIS surface is composed by a metallic patch lying on the PCB, enclosed by the metallic walls of the cavity; b, f) Inset of the lumped LC model; c,g) HFSS simulation configuration; d,h) comparison between HFSS simulations and lumped model.

### 2.8.2 Experimental setup and measurements

First, an idle experiment is realized using a standard waveguide to coaxial transition in order to obtain a rectangular cavity resonator with dimensions width  $a = 23$  mm, length  $d = 33$  mm and height  $b = 10$  mm, as can be seen Figure 2.49. The measured resonance frequency is 7.9 GHz and the extracted unloaded Q-factor, at this frequency is  $Q_0 = 710$ . The measured  $S_{11}$  parameters as well as the E field distribution inside the cavity resonator are also presented in Figure 2.49.

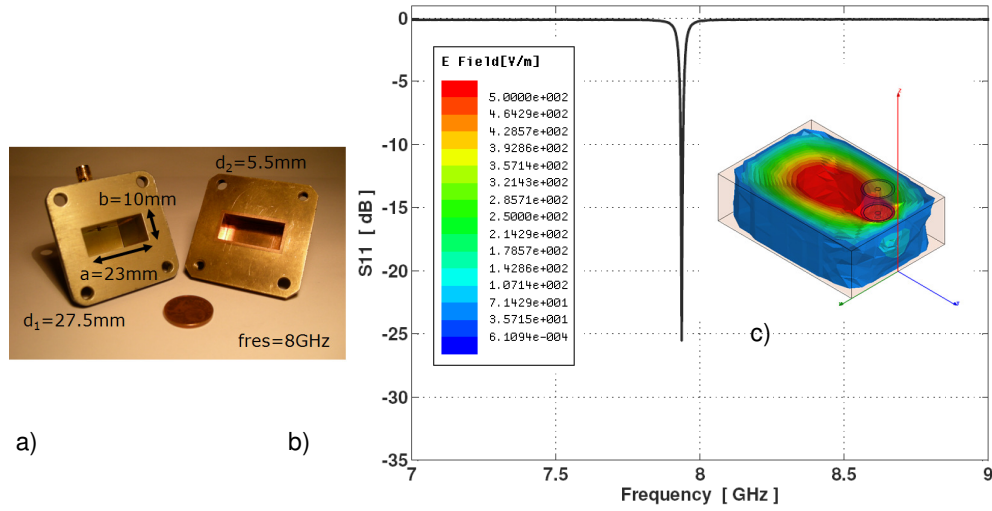


Fig. 2.49.: a) Configuration of a rectangular waveguide cavity in the idle case: full metal cavity, b) measured reflection coefficient,  $S_{11}$  demonstrating a frequency of resonance of 7.9 GHz and c) E field distribution of the  $TE_{101}$  mode of resonance.

Secondly, two prototypes shown Figure 2.50 and Figure 2.51 were realized using standard PCB technology (RO4010 substrate from Rogers Co.) with a dielectric permittivity  $\epsilon_r = 10.2$ ,  $t_{HIS1} = 0.635$  mm,  $t_{HIS2} = 2.5$  mm,  $\tan \delta = 0.0035$  and copper thickness  $17 \mu\text{m}$ .

The first prototype implements the first HIS design and fills the idle cavity resonator with two adjacent HIS walls disposed as in Figure 2.50, covering a width and a length of the idle cavity. The HIS surface is composed by rows of a rectangular metallic patch ( $W_{HIS1} = 4.85$  mm,  $L_{HIS1} = 9.8$  mm) spaced by a slot of  $100 \mu\text{m}$  from the top and bottom of the cavity walls, as can be seen in Figure 2.50. The measured resonance frequency is 4 GHz and the extracted unloaded Q-factor, at this frequency is  $Q_0 = 205$ . The reduction of  $Q_0$  is due to the implementation of the PMC boundary condition by the HIS made of resonant elements, which used at resonance generate additional dielectric and Ohmic losses. The measured  $S_{11}$  parameters as well as the E field distribution inside the cavity resonator are also presented in Figure 2.50.

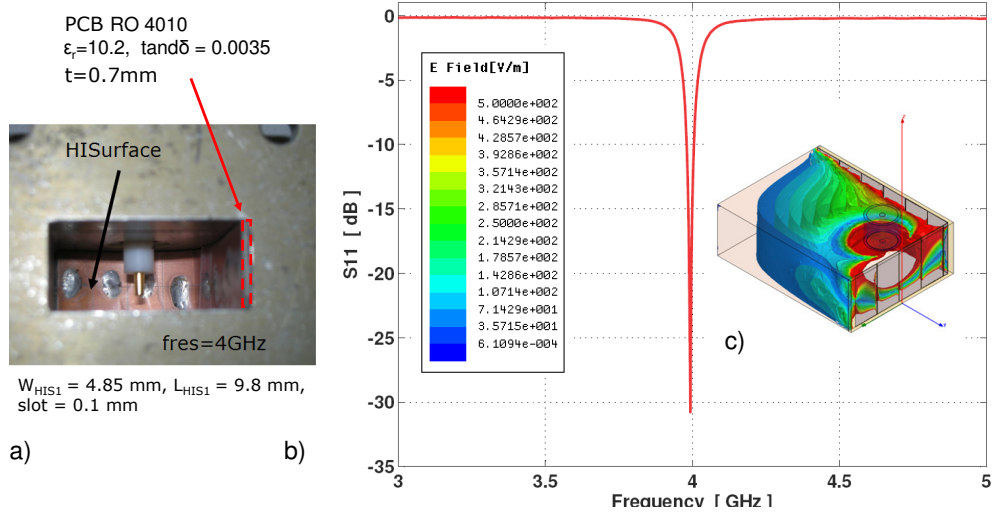


Fig. 2.50.: a) Configuration of a rectangular waveguide cavity with two sidewalls covered by HIS 1, b) measured reflection coefficient,  $S_{11}$  demonstrating a frequency of resonance of 4 GHz and c) E field distribution of the  $\text{TE}_{101}$  mode of resonance.

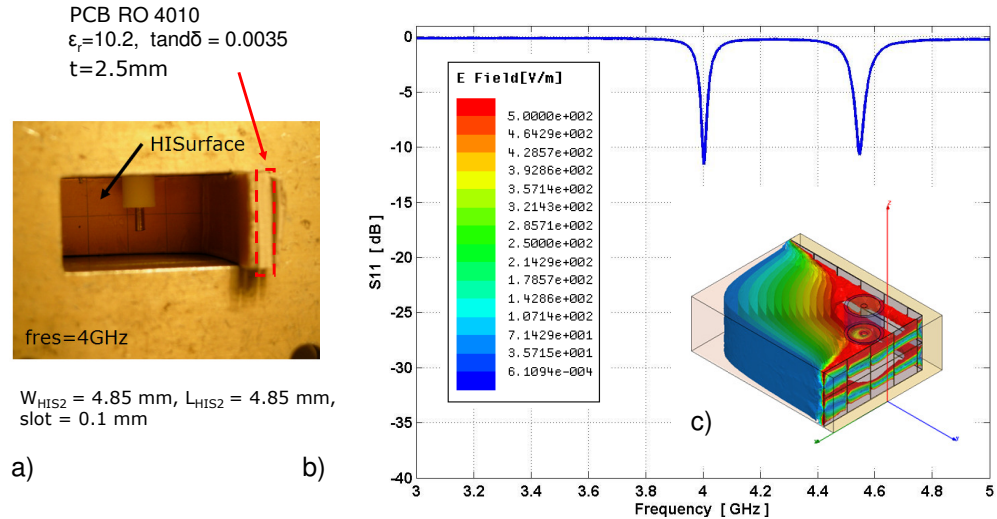


Fig. 2.51.: a) Configuration of a rectangular waveguide cavity with two sidewalls covered by HIS 2, b) measured reflection coefficient,  $S_{11}$  demonstrating a frequency of resonance of 4 GHz but presenting additional spurious resonances and c) E field distribution of the  $\text{TE}_{101}$  mode of resonance.

The second prototype implements the second HIS design and fills also the idle cavity resonator with two adjacent HIS walls disposed as in Figure 2.51, covering a width and a length of the idle cavity. The HIS surface is composed by rows of two rectangular metallic patch ( $W_{HIS2} = 4.85$  mm,  $L_{HIS2} = 4.85$  mm) spaced by a slot of  $100\text{ }\mu\text{m}$  from the top and bottom of the cavity walls, as can be seen in Figure 2.50. The measured resonance frequency is 4 GHz and the extracted unloaded Q-factor, at this frequency is  $Q_0 = 200$ . The reduction of  $Q_0$  is due to the implementation of the PMC boundary condition by the HIS made of resonant elements, which used at resonance generate additional dielectric and Ohmic losses. The measured  $S_{11}$  parameters as well as the E field distribution inside the cavity resonator are also presented in Figure 2.51. Spurious resonant modes are observed which can be generated by additional resonance between the metallic patches.

The new miniaturization concept for cavity resonators demonstrates that there is no need for a cavity resonator to be half a wavelength long and width if boundary conditions are adequately placed, preserving the field distribution of the resonant mode. Nevertheless, the insertion of the high impedance surfaces is also reducing the Q-factor and therefore a trade off between Q-factor performance and compactness should be considered. Measurements fully validate the theoretical predictions i.e. the resonance frequency of a rectangular cavity is shifted from 8 GHz to 4 GHz when two of its sidewalls are covered with the HIS designed for 4 GHz. This concept could be broadened towards the investigation of different other kinds of symmetries, which in turn could increase further the level of miniaturization. However, the integration of such a concept in MEMS technologies was not further investigated within this thesis. Alternative ways to miniaturize cavity resonators are derived from the investigation, that follows on tuning and trimming concepts.

## 2.9 CONCLUSIONS

Precise measurements are essential to characterize RF devices and to verify the accuracy of numerical simulations. To increase the accuracy of the measurements, conventional calibration techniques are used, most commonly the TLR. Three parameters are fundamentally needed for characterization: the unloaded Q-factor, the coupling coefficient and the unloaded resonance frequency, the other parameters can be derived. Actually, the Q-factor is largely affecting the coupling coefficient. Large variations between simulations and measurements in terms of Q-factors are observed. The differences between simulations and measurements observed are explained principally by the additional resistive losses at the junction between the metallized PCB and the cavity resonator, which is considered ideal in the HFSS simulations. Additional sources of loss, which are not enough accounted by the simulations (Ohmic, dielectric and radiative losses) could be accounted by the introduction of the normalized coupling coefficient,  $\kappa_{norm.}$ . Otherwise, this can be corrected by adapting the characteristics of the dielectric substrates, the metallic conductivity and junction resistivity. The characteristics are isolated during specific characterization measurements, while radiation losses are never easy to isolate. Alternatively, the normalization of the coupling coefficient by the ratio measured to simulated Q-factors is providing very accurate results and is our preferred method.

A large sample of cavity resonators provided with different feeds were analyzed and synthetically presented providing the coupling coefficient as a function of a physical parameter, as required prior to any use in any circuit. The characterization of metallic rectangular cavity resonators and various coupling was realized to assess the methodology and to develop the modelling techniques used to design the cylindrical integrated cavity resonators and the integrated resonators in the following chapters. The feeding used in every case is driven by the technology of the implementation. Nevertheless, the concepts of power exchange through the feeding and impedance affected by the location of the feeding are common.

Furthermore, a very high performance integrated cavity resonator for 60 and 75 GHz was realized. These resonators demonstrate typically very high unloaded  $Q$ -factors, respectively  $Q_0 = 117$  and  $Q_0 = 91$  at 60 GHz and 75 GHz, in comparison with integrated inductors for low cost (Bi)CMOS or Bipolar technologies, typically characterized by  $Q$ -factors, limited to around 30 below 10 GHz and decreasing with increasing frequency. Additionally, the cylindrical cavity resonators are fully integrated in IMEC's MCM-D integration platform and no additional manufacturing is required, given the available interconnection TSVs. The overall dimensions of the resonators are respectively  $1.7 \times 1.7 \times 0.1 \text{ mm}^3$  and  $1.5 \times 1.5 \times 0.1 \text{ mm}^3$  at 60 GHz and 75 GHz, which represents an effectively low cost solution as the size on wafer is reduced due to the high permittivity substrate.

The coupling mechanism is affecting the loaded resonance frequency, hiding the unloaded resonance frequency. This is a direct consequence of the energy stored in the feeding, which can be modelled by an additional shunt susceptance or in series reactance. As a consequence, the unloaded resonance frequency needs to be independently extracted. This is realized considering the detuning trend of a set of resonators, characterized by different coupling coefficients. Considering the extracted unloaded resonance frequency, lumped component models are realized. The RF response of the loaded cavity resonators as a function of physical parameters of the coupling structure is modelled. The detuning by the feeding could not be neglected and its impact on the complex impedance is required e.g. in oscillators design. The coupling coefficient is evaluated analytically by computing the energy transfer and considering the impedance governed by the cavity's fields distribution. Practically, both transfer of energy and impedance are defining together the coupling coefficient. Using lumped components models of the loaded cavity resonators saves time otherwise consumed with heavy full wave HFSS simulations. These further improve the understanding of the physical behavior of the loaded cavity resonators while helping to identify measurement and manufacture errors.

Finally, a new miniaturization concept for cavity resonators was demonstrated. It is demonstrated that there is no need for a cavity resonator to be half a wavelength long and width if boundary conditions are adequately placed, preserving the field distribution of the resonant mode. More precisely and for the  $TE_{101}$  mode of resonance, placing PMC boundary conditions at quarter wavelength long and width provides the same field distribution as in the idle case where PEC boundary conditions were placed half a wavelength away. The practical implementation of the PMC boundary condition is realized using High Impedance Surfaces (HIS) designed using either a lumped element approach or full-wave simulations. Each element of the HIS is composed by metallic



patches lying on PCB and enclosed by the metallic walls of the cavity resonator. Nevertheless, the insertion of the high impedance surfaces is also reducing the Q-factor and therefore a trade off between Q-factor performance and compactness should be considered. Measurements fully validate the theoretical predictions i.e. the resonance frequency of a rectangular cavity is shifted from 8 GHz to 4 GHz when two of its side-walls are covered with the HIS designed for 4 GHz. This demonstrates equivalently that a cavity resonator with HIS operating at 4 GHz has half the length and width of a standard metallic cavity resonator and its volume is divided by four. However, the integration of this concept with microfabrication techniques was not investigated in this thesis. Nevertheless, alternative ways to miniaturize cavity resonators are derived from the tuning investigation, which will be presented in Chapter 3 and for which the integration was realized.



## CHAPTER 3

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# INTERNAL VOLUMES OF PERTURBATION FOR TUNING CAVITY RESONATORS

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### 3.1 INTRODUCTION

Besides fixed frequency cavity resonators, there is a need for tunable cavity resonators for use in tunable filters [Park 08]. Tunable and step trimmed cavity resonators are also considered to reduce the need and cost of a manual trimming and to allow an electrical tuning during operation [Tilmans 07]. The need for tunable cavity resonators is further clearly recognized [White 05, Park 09, Yan 07, Kim 99, Song 09], as presented in Chapter 1.

In this chapter, we investigate the tuning performance induced by an internal volume of perturbation. The internal volume of perturbation is either a Central Post Perturbation, CPP or a Side Annular Perturbation, SAP. The CPP perturbs dominantly the electric field distribution, which is maximal in magnitude in the center of the cavity resonator, affecting the resonance frequency of the cavity resonator, by reducing it. The lateral dimensions and the height of the CPP determine the shift in frequency. By making the lateral dimensions and the height variable, the resonance frequency can be tuned. The SAP perturbs dominantly the magnetic field distribution, which is maximal in magnitude at the walls of the cavity resonator, affecting the resonance frequency of the cavity resonator, by increasing it. The tuning mechanism is fully described using the small perturbation theory applied to the volume incorporated in the cavity resonator. The impact of the insertion of such perturbations is assessed by eigen mode HFSS simulations and measurements of 72 different cavity resonators prototypes, tuned from 1 to 6 GHz from an idle resonance frequency of 4 GHz. A maximal frequency shift of 3 GHz and 2 GHz is measured for respectively the resonator with

CPP and SAP. This is resulting in a total available tuning range of 125%. The analytical evaluation of the resonance frequency as well as the evaluation of the Q-factor are compared to HFSS simulations and measurements. Additionally, the CPP case could be used for miniaturization purposes, as cavity resonators with a resonance frequency as low as 1 GHz were measured, derived from an idle cavity resonator at 4 GHz.

Finally, we focus on the integration of a tunable cavity resonator at 60 GHz. This implementation can be seen as a variable-sized MEMS-based Faraday cage [Dancila 10], used to tune the resonance frequency of a silicon micromachined cavity resonator. The Faraday cage is defined by a 2D array of electrostatically controllable metal cantilevers. The array of cantilevers effectively shields an enclosed (air) volume forming the Faraday cage, located inside the micromachined cavity resonator. The Faraday cage perturbs the electromagnetic field distribution inside the cavity resonator and provides a means to electrically tune the resonance frequency of the resonator. The cavity resonator is micromachined using bulk KOH etching in silicon and assembled in a low temperature die-to-wafer soldering process. The electrical feeding is realized in thin-film metallizations and the resonator is flip chip assembled. This implementation offers a fully integrated solution and distinguishes itself by enclosing the MEMS inside the cavity, thereby conveniently using the cavity at the same time as a 0-level package for the MEMS.

### 3.2 INTERNAL PERTURBATION TUNING MECHANISM

The link between the size of the cavity, the mode of resonance and the resonance frequency were given in Section 2.4. It was also mentioned that the smallest cavity resonator is half a wavelength long and the highest possible unloaded Q-factor is provided by a square cavity ( $a = d$ ), as defined in Figure 2.8.

A common tuning technique is changing the boundary conditions by displacing a moving wall inside a cavity resonator. This reduces the effective length of the cavity and thus the resonance frequency, although this method was not preferred due to the limitations of the available technology. Another technique consists in connecting posts located in the cavity resonator by means of MEMS. Some preliminary design and modelling work was done within this thesis, but no further investigation was undertaken. Instead, we investigated an implementation with relatively low excursion MEMS, in comparison with the total height of the cavity. The MEMS define controllable volumes located inside the cavity resonator. The MEMS are adapted from the Electrostatic Fringing Field Actuators (EFFA) by [Rottenberg 07]. The small perturbation theory is used to study the resonance behavior of these perturbed cavity resonators. The resonance frequency is computed by evaluating the ratio between the electric and magnetic energy stored in the cavity resonator, reported to the energy un-stored in the perturbed volume.

We investigated therefore the concept of a stand-alone variable perturbation, located inside the cavity resonator, actuated from outside the cavity resonator accordingly to an initial concept stated in [Tilmans 07], that we further improved. The tuning is of mechanical means and therefore a physical variation is created inside the cavity

resonator which in turn is tuning the resonance frequency. Integration being the final scope, the variable volume should be foldable and un-foldable within the cavity resonator without any external force.

Among the main contributions of this thesis, we certainly point out the investigation of the tunability by using perturbation volumes, which heights and lateral dimensions are made variable. The investigation concerns both the tunability of the resonance frequency and the evaluation of the unloaded Q-factor resulting from the insertion of such perturbations. Finally, optimal perturbation shapes are derived in order to maximize the tuning range given the relatively small excursion of the given MEMS. A schematic view of the proposed concept is depicted in Figure 3.1.

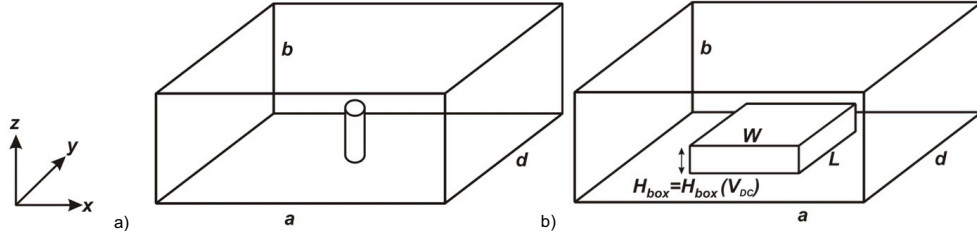


Fig. 3.1.: Rectangular cavity resonator (dimensions  $a \times d \times b$ ) with enclosed a) thin post and b) rectangular perturbation (dimensions  $W \times L \times H_{box}$ ).

In Figure 3.1-b) a rectangular cavity (with dimensions  $a \times d \times b$ ) is shown, in which a rectangular post (with dimensions  $W \times L \times H_{box}$ ) is placed at the bottom-center. The rectangular post perturbs the electromagnetic field distribution thereby affecting the resonance frequency of the cavity resonator. The size of the rectangular post determines the shift in frequency. By making the height of the perturbation box,  $H_{box}$  variable, the resonance frequency can be tuned. The perturbation is located in the middle of the cavity, in order to maximize the tuning as the intensity of the electrical field is maximal in the center of a  $TE_{101}$  mode cavity.

The perturbation technique is commonly applied to the analysis of the shift of the resonance frequency as a consequence of the insertion of a (thin) tuning post, with volume  $\Delta V$ , located exactly in the center of a cavity resonator of volume  $V_o = abd$ , as shown in 3.1-a). The relative frequency shift is approximatively given by [Pozar 05]:

$$\frac{f - f_o}{f} = -2 \frac{\Delta V}{V_o} \quad (3.1)$$

where  $f$  and  $f_o$  denote the perturbed and unperturbed resonance frequency, respectively. The approximation (3.1) is inherited from the era of RF waveguides with a

screw inserted inside a cavity resonator, for tuning purposes. The tuning evaluation is based on the hypothesis that the field distribution is only slightly disturbed by the perturbation as to approximate the field distribution by the unperturbed one. Even though we are aware of the limitations of the approximation limited to a thin post located in the center of the resonator, we propose to use it in the case of a rectangular perturbation located in the center of the cavity, as presented in the Figure 3.1-b). It can be stated that the relative frequency shift is a linear function of  $H_{box}$ , following (3.1).

In a second approach, the field distribution of the  $TE_{101}$  mode (2.60) are considered, instead of the fields homogeneity, as in (3.1). The result is more realistic, as the field distribution, as it is, is taken in account. The resonance frequency is computed by evaluating the electrical energy  $W_e$  and magnetic energy  $W_m$  stored in both the perturbed and initial volume of the cavity, as follows:

$$W_e = \frac{\varepsilon}{4} \int_v |\vec{E}|^2 dV \quad (3.2)$$

$$W_m = \frac{\mu}{4} \int_v |\vec{H}|^2 dV \quad (3.3)$$

where  $E$  and  $H$  are the unperturbed electric and magnetic fields, respectively defined in (2.60). The frequency shift is given by the fraction of the stored energy in the perturbed volume related to the stored energy in the initial non perturbed cavity [Pozar 05]:

$$\frac{f - f_0}{f_0} \cong \frac{\int_{\Delta V} (\mu |\vec{H}|^2 - \varepsilon |\vec{E}|^2) dV}{\int_{V_0} (\mu |\vec{H}|^2 + \varepsilon |\vec{E}|^2) dV} \quad (3.4)$$

or using (3.2) in a synthetic form:

$$\frac{f - f_0}{f_0} \cong \frac{\Delta W_m - \Delta W_e}{W_m + W_e} \quad (3.5)$$

Both approximative method (3.1) and integral method (3.4) are exposed in the case of a rectangular cavity resonator (dimensions  $a = d = 3844 \mu m$ ,  $b = 440 \mu m$ ) with enclosed rectangular metallic box (dimensions  $W = L = 1900 \mu m$  and  $H_{box}$  is made variable). Those dimensions correspond to a  $TE_{101}$  mode of a rectangular cavity resonator with a resonance frequency of 55 GHz. We will see later in Section 3.3 that this corresponds to a trapezoidal cavity resonator of 60 GHz. The integral method evaluates lower the frequency shift as the HFSS simulation does. This is because the fields are initially considered unperturbed, something which is not the case in the HFSS simulations, where the real fields distribution are computed for a given perturbation, see Figure 3.2. There is only a 0.5% difference between the integral method and the HFSS simulations for a perturbation height larger than 12% of the cavity height ( $50/440 \mu m$ ), see Figure 3.2.

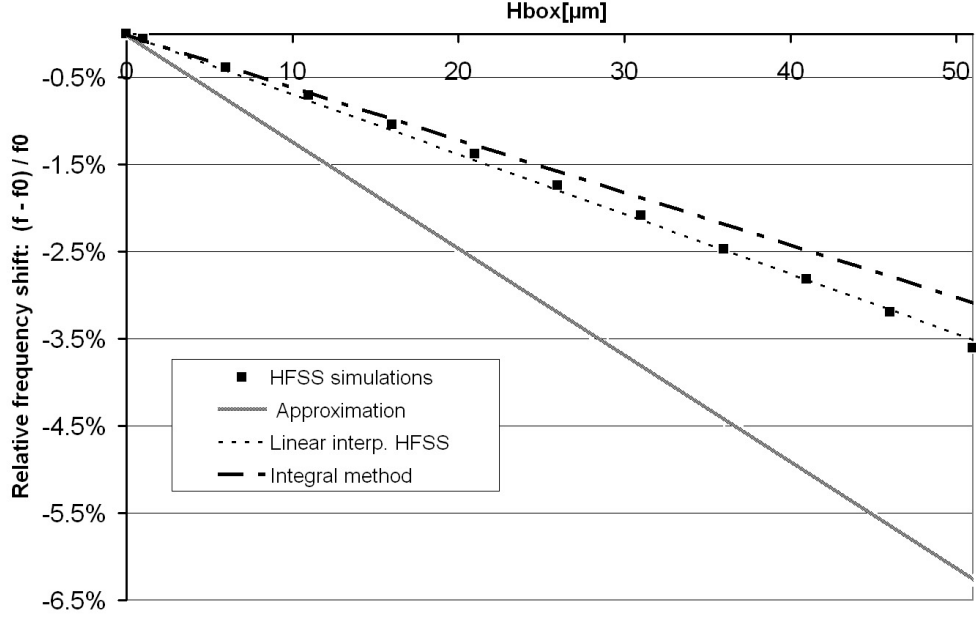


Fig. 3.2.: Comparison between three methods to evaluate the tuning resulting from the perturbation by a volume located inside the cavity resonator: the approximation (3.1), the integral method (3.4) and HFSS simulations. The simulations apply for a rectangular cavity resonator with dimensions:  $a = d = 3844 \mu m$  and  $b = 440 \mu m$ , and inner rectangular post dimensions  $W = L = 1900 \mu m$ , with variation of the height  $H_{box}$  from 0 to  $50 \mu m$ .

### 3.2.1 Prototype at 4 GHz

This section investigates further the implications of the perturbation theory on the resonance behavior of a cavity resonator incorporating a perturbation volume located internally to the cavity resonator. Studying (3.4) and considering the inhomogeneous field distribution (2.60) of the  $TE_{101}$  mode of resonance, we identify two cases. The perturbation volume located in the center of the cavity resonator is called the Central Post Perturbation (CPP), presented in Figure 3.3-a) with a corresponding cross-section in Figure 3.3-b). The perturbation volume filling the surroundings of the center is called the Side Annular Perturbation (SAP), presented in Figure 3.3-c) with a corresponding cross-section in Figure 3.3-d). For symmetry reasons, we consider the perturbation (post or hole) having the same lateral dimensions  $W = L$  (square section) and we locate the perturbation in the center of the cavity resonator. As a consequence, only two parameters can be made variable, the height,  $H_{box}$  and the lateral dimensions of the perturbation volume. The variation of these parameters and the evaluation of the impact on the resonance frequency and Q-factor are fully investigated in this section. The impact of different lateral dimensions and heights of both CPP and SAP volumes is investigated by evaluating analytically the resonance frequency of the perturbed

resonator using the small perturbation theory [Collin 66]. The idle cavity resonator prototype of Chapter 2 is reused. Additionally, a set of 72 cavity resonators incorporating both CPP and SAP of different and fixed lateral dimensions and heights are simulated with HFSS, realized in aluminum and measured. These are assessing the limitations of the small perturbation theory and demonstrate the large tuning capabilities by this approach.

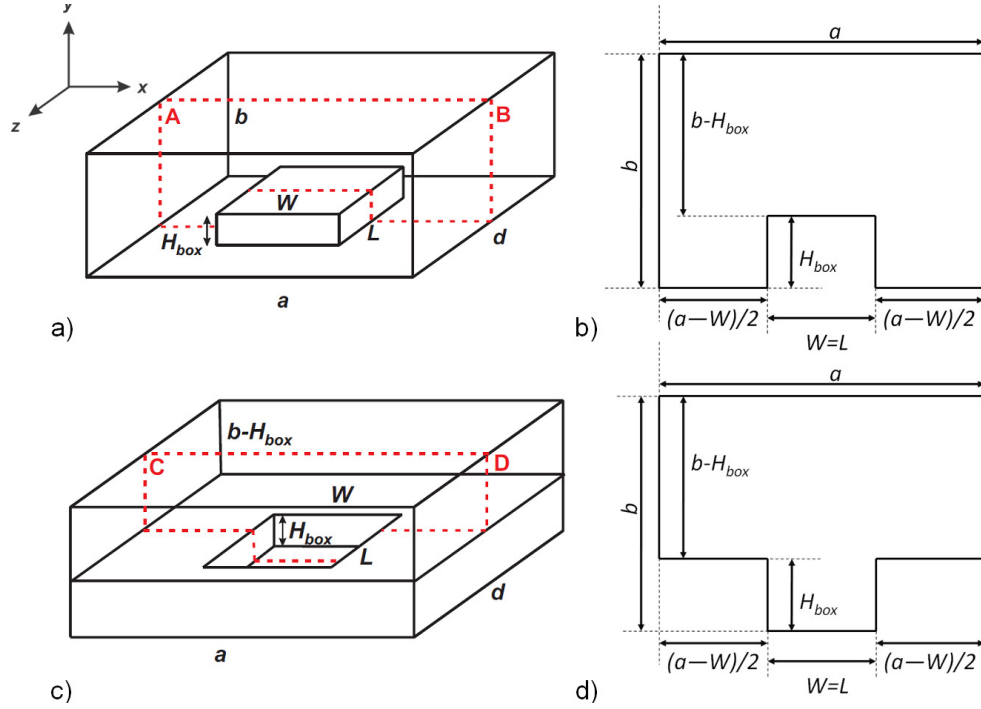


Fig. 3.3.: a) Rectangular cavity resonator (dimensions  $a \times d \times b$ ) with enclosed central post perturbation (CPP, dimensions  $W \times L \times H_{box}$ ), b) cross section A-B, c) enclosed side annular perturbation (SAP's hole dimensions  $W \times L \times H_{box}$ ) and d) cross section C-D.

The 72 prototypes, address both Central Post Perturbation (CPP) and Side Annular Perturbation (SAP) in combinations of the two parameters mentioned here above are described in the Table B.1 from Appendix B. Additionally the idle case corresponds to the cavity resonator previously discussed in Chapter 2 ( $a = d = 53$  mm,  $b = 25$  mm) with an idle resonance frequency at 4 GHz. The arrays of cavity resonators arranged in a matrix of 6 by 6 cavity resonators is presented in Figure 3.4. The Figure 3.4 presents a perspective view of the array of cavities with CPPs and SAPs and the overview of the final implementation. The lateral dimensions of the matrix is 400 mm by 400 mm. Each cavity resonator incorporates a different internal volume of perturbation either CPP or SAP. The variation of the height is ranging from 5 to 24 mm ( $b = 25$  mm) in 5 steps of 5 mm (i.e. 5, 10, 15, 20 and 24 mm). The variation of the lateral dimensions



is ranging from 5 to 40 mm ( $a = d = 53$  mm) in 7 steps (i.e. 5, 10, 15, 20, 25, 30 and 40 mm). The analytical evaluation of the resonance frequency, the HFSS simulations and measurements are further presented.

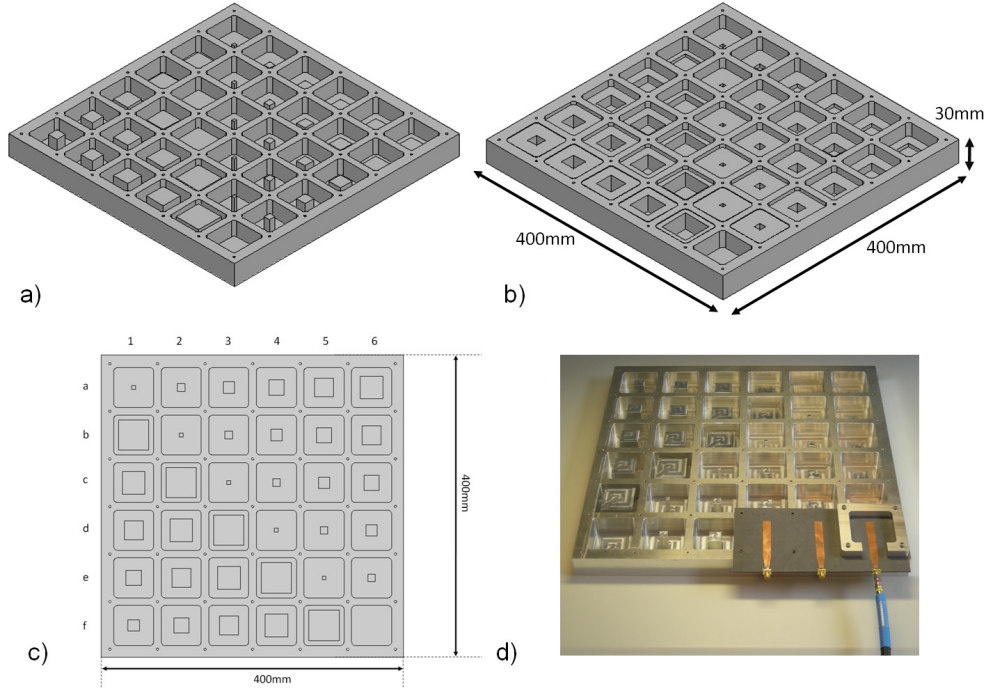


Fig. 3.4.: Array of cavity resonators with a) Central Post Perturbation (CPP); b) Side Annular Perturbation (SAP); c) Names and locations of the samples referring to the Table B.1 from Appendix B and d) photo of the array and assembled resonator, ready for measurement.

### 3.2.2 Analytical evaluation of the resonance frequency

The stored electrical energy,  $W_e$  and magnetic energy,  $W_m$  are evaluated in the volume of the Central Post Perturbation (CPP), as follows.

$$W_e = \frac{\varepsilon}{4} \int_{a/2-W/2}^{a/2+W/2} \int_{d/2-W/2}^{d/2+W/2} \int_0^{H_{box}} |\vec{E}|^2 dx dz dy \quad (3.6)$$

$$W_m = \frac{\mu}{4} \int_{a/2-W/2}^{a/2+W/2} \int_{d/2-W/2}^{d/2+W/2} \int_0^{H_{box}} |\vec{H}|^2 dx dz dy \quad (3.7)$$

where the fields are considered unperturbed, as in the idle case (2.60). The results of this calculation are presented in Figure 3.5. The stored energies are monotonously

growing with growing lateral dimensions. The stored electrical energy is always higher than the magnetic stored energy, as the perturbation is centrally located in the cavity resonator, where the electric field is dominant. At the maximum perturbation limit, when the lateral dimensions equal the cavity dimensions ( $a = d = W = L$ ) both stored energies are equal. The impact of the height is only a scale factor, as the fields are not a direct function of the height, in the  $TE_{101}$  mode of resonance.

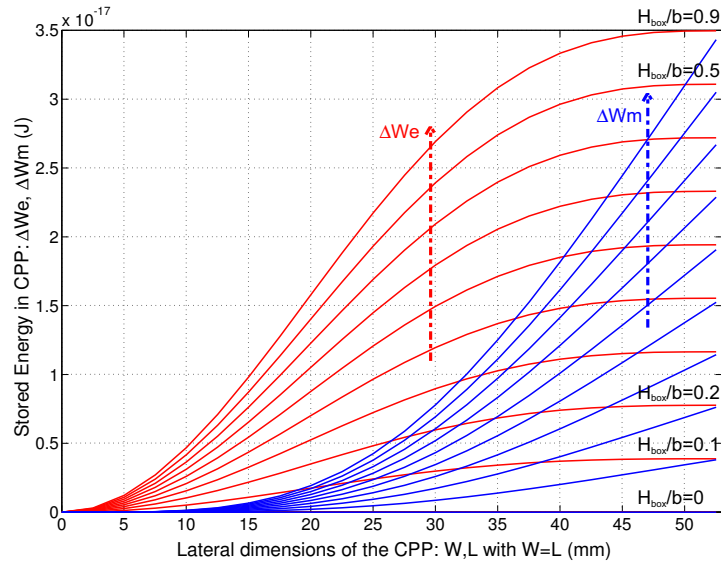


Fig. 3.5.: Stored Electric Energy ( $W_e$  red) and Magnetic Energy ( $W_m$  blue) in the CPP's volume.

The evaluation of the difference  $W_m - W_e$  of (3.5) between the stored energies inside the perturbation volume is further presented in Figure 3.6. As the electrical energy is always higher than the magnetic energy, the difference is negative. This results further in a negative frequency shift as the denominator of (3.5) is positive. Placing the internal volume of perturbation in the center of the cavity resonator has as consequence that the resonance frequency decreases, as can be seen in Figure 3.7. It is also found, that a maximal tuning range was achieved for CPP that are approximately half the size of the cavity. More precisely, for a CPP with lateral dimension  $W = L \sim 30$  mm. After this value, the magnetic stored energy in the CPP is increasing more rapidly with larger side dimensions, because closer to the edge of the cavity the magnetic field is dominant. Furthermore, the larger the CPP's height, the larger the shift in frequency.

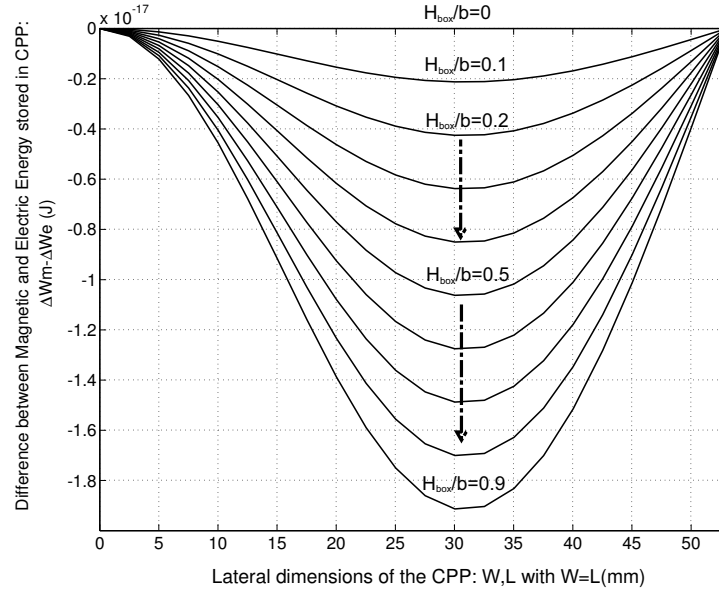


Fig. 3.6.: Difference between the magnetic and electric energy in the CPP.

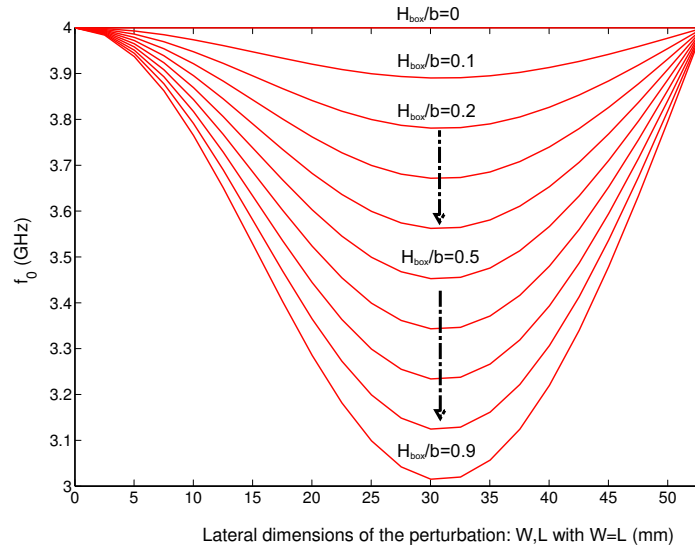


Fig. 3.7.: Frequency of resonance as resulting the insertion of the CPP.

Similarly, in the case of the Side Annular Perturbation (SAP), the electric and magnetic energy stored in the SAP are evaluated considering the cavity in Figure 3.3-c)-d) as composed of two parts: an upper part above the SAP of height  $H_{box}$  and a lower part in the hole of the SAP. Therefore the stored energies are evaluated as follows:

$$W_{e,m_{SAP}} = W_{e,m} - W_{e,m_{UP}} - W_{e,m_{CPP}} \quad (3.8)$$

where  $W_{e,m}$  is the electric, magnetic energy stored in the cavity, respectively (3.2),  $W_{e,m_{CPP}}$  is given by (3.6) and  $W_{e,m_{UP}}$  is calculated, as follows:

$$W_{e_{UP}} = \frac{\varepsilon}{4} \int_0^a \int_0^d \int_{H_{box}} |\vec{E}|^2 dx dz dy \quad (3.9)$$

$$W_{m_{UP}} = \frac{\mu}{4} \int_0^a \int_0^d \int_{H_{box}} |\vec{H}|^2 dx dz dy \quad (3.10)$$

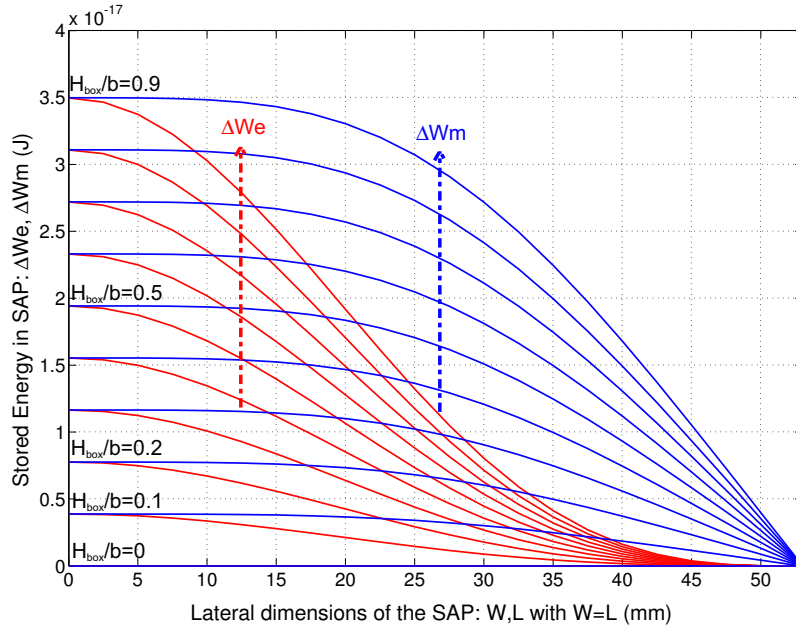


Fig. 3.8.: Stored Electrical Energy ( $W_e$  red) and Magnetic Energy ( $W_m$  blue) in the SAP.

The results of the calculation (3.8) are presented in Figure 3.8. The stored energies in the SAP are monotonously decreasing with the growing lateral dimensions of the SAP. The first idle state is when the SAP has no lateral dimensions,  $W = L = 0$ . In this case, the stored electrical energy,  $W_e$  and magnetic energy,  $W_m$  are equal. Remarkable is also to see that the energy stored in the SAP increases linearly with  $H_{box}$  as the fields, for the  $TE_{101}$  are not a direct function of the height but the height contributes to the volume of integration. The second idle state is when the SAP's dimensions are equal to the cavity's dimensions ( $a = d = W = L$ ) and then there is no stored energy in the SAP, the cavity is the hole. In the other cases, the magnetic stored energy is always higher than the stored electrical energy, as the perturbation located around the center of the cavity resonator, where the magnetic field is dominant.

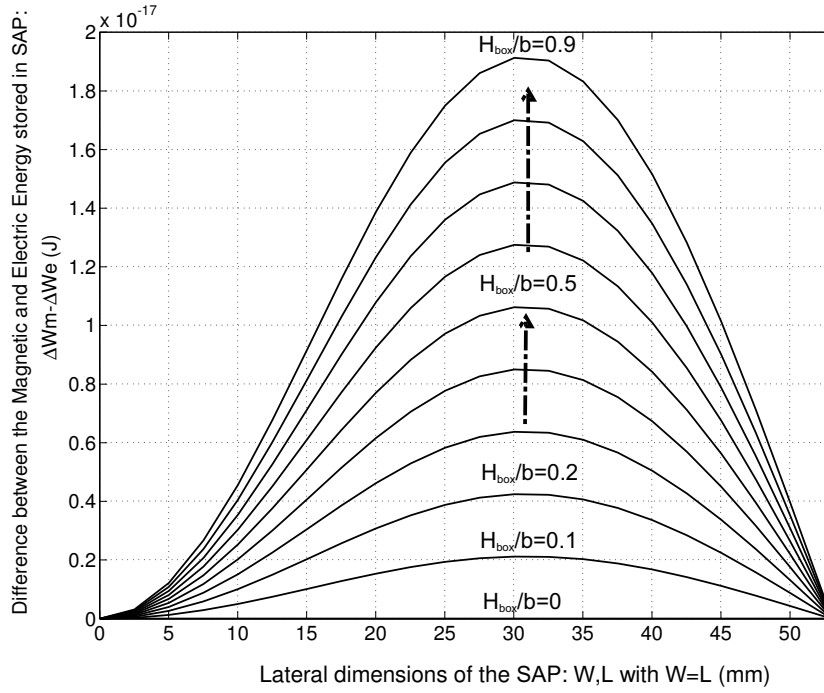


Fig. 3.9.: Difference between the magnetic and electric energy in the SAP.

The evaluation of  $W_m - W_e$  (3.5) between the stored energies inside the SAP volume is presented in Figure 3.9. As the magnetic energy is always higher than the electric one, the difference is positive, which results further in a positive frequency shift as the denominator (3.5) is positive. As can be seen in Figure 3.10, the SAP is increasing the resonance frequency. The maximal tuning range was achieved for the SAP of approximately half the size of the cavity ( $W = L \sim 30$  mm), similar result as in the CPP case. Before this, the stored electrical energy in the SAP decreases faster with

larger hole lateral dimensions, as the center of the cavity where the electric field is dominant, is let open. After the maximal value, the magnetic energy stored decreases as the hole is reducing the SAP.

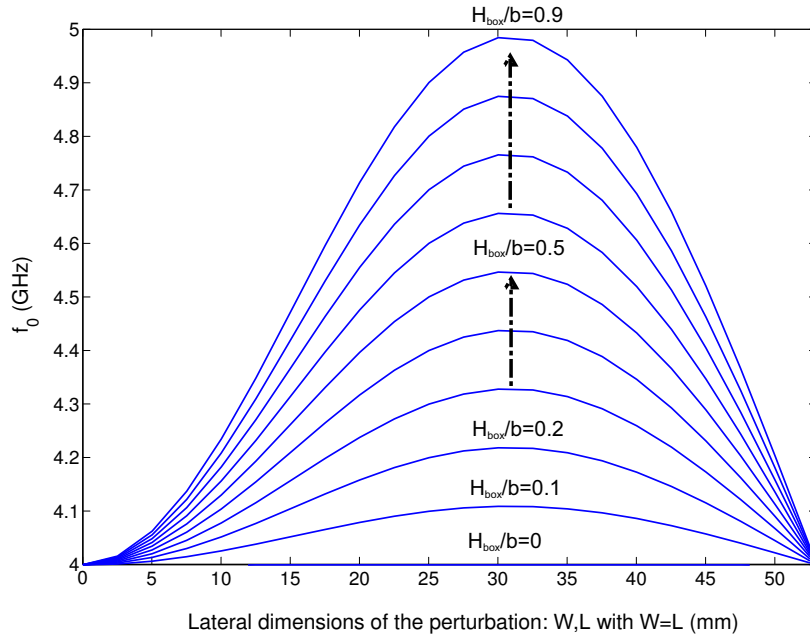


Fig. 3.10.: Frequency of resonance as resulting the insertion of the SAP.

### 3.2.3 Resonance frequency: HFSS simulations and measurements

The simulation set-up is presented in Figure 3.11. It is composed of a cavity resonator with either CPP or SAP located in its center. For obvious symmetry reasons, only one quarter of the cavity with perturbation is simulated. Each cavity resonator incorporates a different internal volume of perturbation either CPP or SAP. The variation of the height is ranging from 5 to 24 mm ( $b = 25$  mm) in 5 steps of 5 mm (i.e. 5, 10, 15, 20 and 24 mm). The variation of the lateral dimensions is ranging from 1 to 51 mm ( $a = d = 53$  mm) in 26 steps of 2 mm. The resonance frequency and Q-factor of the fundamental mode are evaluated with HFSS eigen mode simulations. The prototypes of the matrix with cavity resonators with perturbation, presented in Figure 3.4, sweep the height,  $H_{box}$  from 5 to 24 mm ( $b = 25$  mm) in 5 steps of 5 mm (i.e. 5, 10, 15, 20 and 24 mm). The lateral dimensions sweep from 5 to 40 mm ( $a = d = 53$  mm) in 7 steps (i.e. 5, 10, 15, 20, 25, 30 and 40 mm). The measurements were performed

with a microstrip line feed with a unitary fraction diameter of the aperture in the PCB's bottom metallization to the width of the microstrip line,  $o/w = 1$ . Following the measurements, this feeding provides a slightly over coupled feeding, with  $\kappa = 2.3$ . Actually, following the feeds assessment realized in Section 2.4.2, the extracted coupling coefficients are lower than designed, as a consequence of the Q-factor reduction. Therefore, this microstrip line feeding is characterized by the highest coupling among the ones developed in Section 2.4.2.

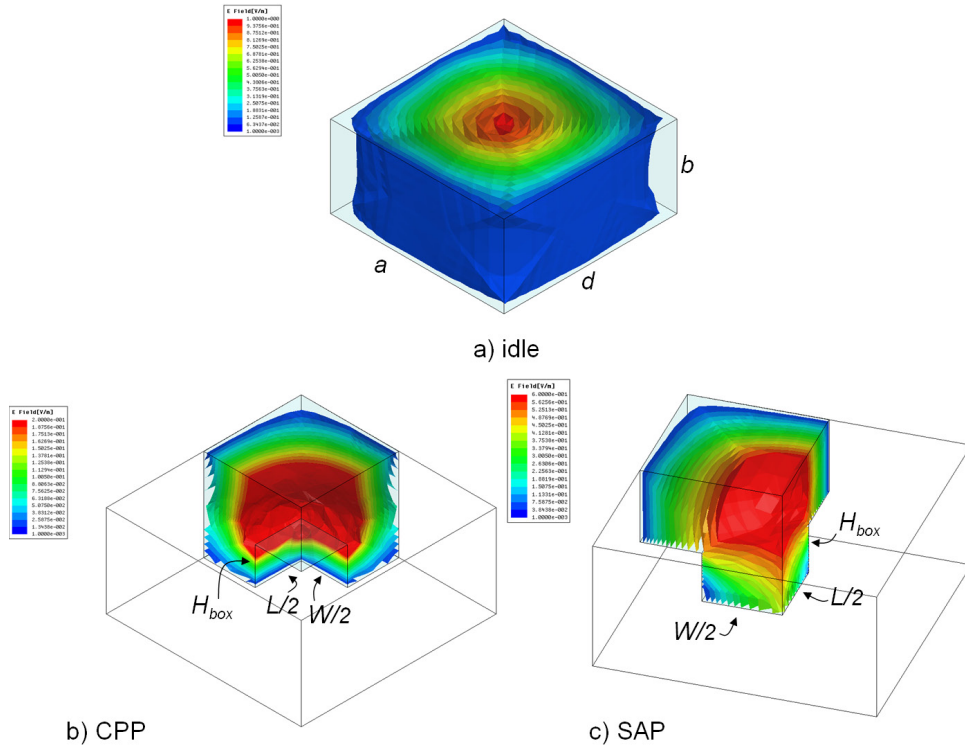


Fig. 3.11.: Simulation set-up in HFSS and E field distribution a) in the idle case with no perturbation, b) with a Central Post Perturbation (CPP) and c) with a Side Annular Perturbation (SAP).

As we can see in Figure 3.12, in the case of the CPP and in Figure 3.13, in the case of the SAP, the eigen mode simulations realized with HFSS are in very good agreement with the measurements. This result is quite remarkable as the measurements are performed on cavity resonators provided with feeding and we know from Section 2.6 that the coupling is detuning the resonance frequency. Nevertheless in the case of very high Q-factor resonators, this detuning is negligible. The resonance frequency was extracted from measurements with the QZERO, as described in Section 2.3.4.

Finally, we compare the HFSS simulations and measurements with the analytical development in Section 3.2.2 and we remark that the behavior and trends are well

similar. Nevertheless, there are some discrepancy that are explained by limitations of the hypothesis which lie at the bases of the analytical modelling. Actually, in the analytical evaluation the fields are considered unperturbed while in HFSS simulations and measurements the field distribution resulting the insertion of the perturbation is relatively different from the idle case where there is no perturbation, as can be seen in Figure 3.11. The consequences are visible when are compared Figure 3.12 with Figure 3.7, for the CPP case and Figure 3.13 with Figure 3.10, for the SAP case.

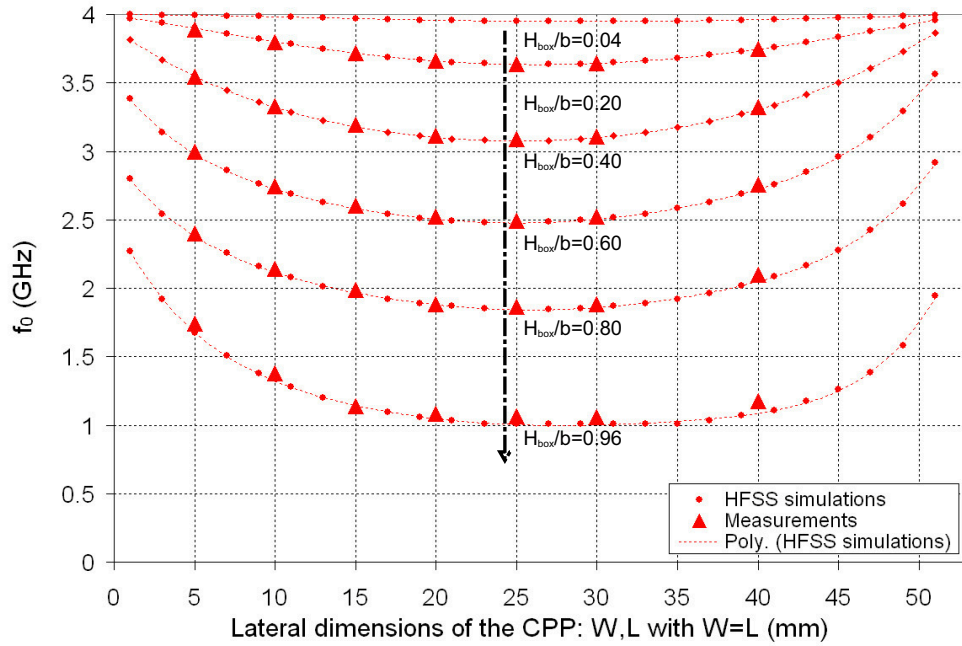


Fig. 3.12.: HFSS simulations and Measurements of the tuning by a Central Post Perturbation (CPP), as a function of both lateral dimensions of the perturbation post and height.

In the CPP case, the impact of the post perturbation is higher than accounted analytically: a maximal frequency shift of 1 GHz is derived from (3.4) while from HFSS simulations and measurement the frequency shift is close to 3 GHz. The post perturbation is closer to the top of the cavity where the E field is the highest, as can be seen in Figure 3.11-a). Instead the E field being straight as considered by the analytical integral method (the field considered are the idle E fields, not perturbed), in reality the E fields points perpendicularly to the post surface towards the top inner face of the cavity resonator. Charges are packed at the top of the post resulting in a potential difference between the top of the cavity and the top of the inner perturbation post.



As a consequence, the evaluation of the stored energy in the perturbation volume, based on the integration of the E field distribution over the perturbation volume, (3.4) is in reality higher as the E field distribution is higher. The HFSS simulations predict better the tuning due to the insertion of a perturbation volume because of fringing fields and edge effects. The higher the field intensity, the higher the magnitude of the integral over the perturbation volume.

This is also visible for small lateral dimensions of the CPP, as the sharp edges of the perturbation bunch together the E fields lines around the perturbation. Nevertheless, the trends are similar as the frequency is monotonously shifted downwards with an increasing CPP's height. The downwards frequency shift is increasing with the perturbation's lateral dimensions, reaching a maximum for CPP's lateral dimensions around the half of the cavity's dimensions i.e. 30 mm for the analytical evaluation and 25-30 mm for the measurements.

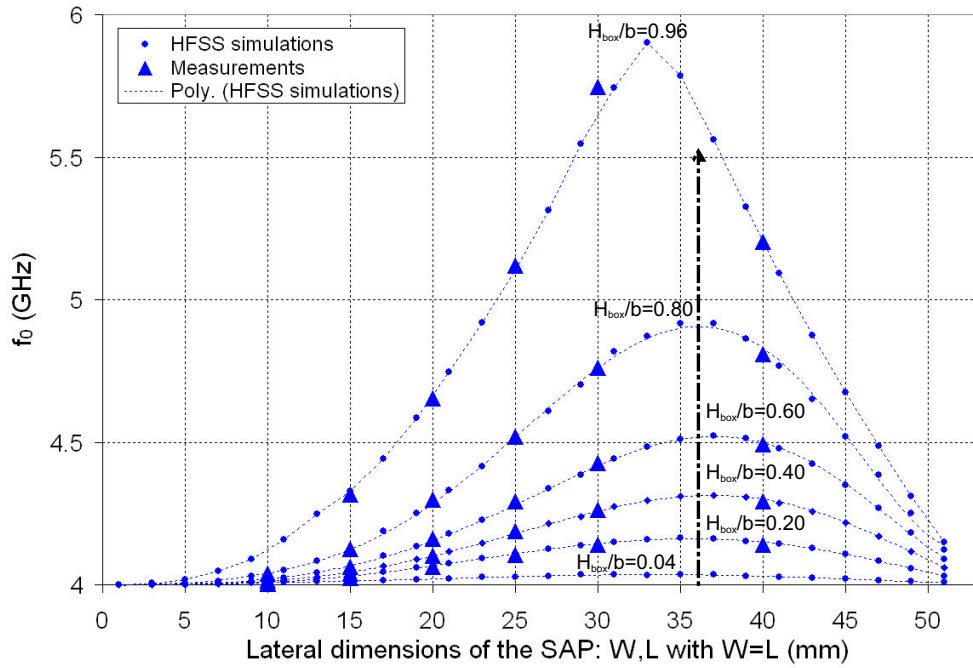


Fig. 3.13.: HFSS simulations and Measurements of the tuning by a Side Annular Perturbation (SAP), as a function of both lateral dimensions of the perturbation hole and height.

In the SAP case, the impact of the internal volume of perturbation is also higher than calculated analytically but smaller than in the CPP case: a maximal frequency shift of 1 GHz is derived from (3.4) while from HFSS simulations and measurement this is

close to 2 GHz. In this case the field perturbation is less pronounced and the difference with the unperturbed field distribution is lower than in the CPP case. This is especially the case for small dimensions of the SAP's lateral dimensions of the centrally located hole, as the volume is shaped by large metallic pieces and the small dimensions of the hole do not allow the E field to engulf or further be perturbed unless dimensions are not negligible in comparison with the cavity's dimensions. In this case also, the sharp edge effect is increasing the E field distribution which serves to evaluate the frequency shift and therefore larger frequency shifts are measured. The higher the field intensity, the higher the magnitude of the integral over the perturbation volume. Nevertheless, the trends are again very similar between the analytical evaluation and the measurements. The frequency is monotonously shifted upwards with increasing SAP's height. The upwards frequency shift is increasing with the perturbation's lateral dimensions, reaching a maximum for SAP's lateral dimensions around the half of the cavity's dimensions i.e. 30 mm for the analytical evaluation, 35 mm for the measurements. The theoretical developments are in good agreement with the measurements, even though the latter show a higher impact on the frequency tuning than expected following the analytical calculation. Nevertheless, they are very accurately simulated with HFSS. This result has the following implications. First, it was demonstrated that both the variation of the height and the variation of the lateral dimensions of the perturbation have a direct impact on the resonance frequency. Second, using only the height variation an optimal shape of the perturbation can be used to perform the highest impact on the frequency tuning of a cavity resonator. Third, using a perturbation volume composed of an array of commutable volumes of binary height (i.e. 0 or  $H_{box}$ ), the frequency could be made tunable by the commutation of selected different parts of the perturbation volume i.e. CPP or SAP following a fractional increase in its whole lateral dimensions,  $W$  and  $L$ .

### 3.2.4 Analytical evaluation of the Q-factor

In this section we perform an analytical evaluation of the Q-factor resulting from the insertion of the perturbation regardless of CPP or SAP. The results are compared to HFSS simulation and to measurements. The measurements correspond to the array of cavity resonators with fixed perturbation previously described in Section 3.2.1. In the case of an air cavity resonator, considering no losses in the filling material, (2.23) results in  $Q_0 = Q_{cond}$  and the unloaded Q-factor is evaluated, as follows [Pozar 05]:

$$Q_0 = \frac{2\omega_0 W_e}{P_c} \quad (3.11)$$

where  $\omega_0 = 2\pi f_0$ ,  $W_e$  is the electric energy stored in the cavity resonator and  $P_c$  is the power dissipated in the cavity walls:

$$P_c = \frac{R_s}{2} \int_{walls} |\vec{H}_t|^2 ds \quad (3.12)$$

with  $H_t$  the tangential magnetic field at the surface of the walls and  $R_s$ :

$$R_s = \sqrt{\frac{\omega\mu_0}{2\sigma}}. \quad (3.13)$$

Using the field distribution (2.60),  $P_c$  is evaluated as follows:

$$\begin{aligned} P_c = \frac{R_s}{2} & \left\{ 2 \int_{y=0}^b \int_{x=0}^a |H_x(z=0)|^2 dx dy \right. \\ & + 2 \int_{z=0}^d \int_{y=0}^b |H_z(x=0)|^2 dy dz \\ & \left. + 2 \int_{z=0}^d \int_{x=0}^a \left[ |H_x(y=0)|^2 + |H_z(y=0)|^2 \right] dx dz \right\} \end{aligned}$$

Additionally, in both CPP and SAP cases power dissipation has to be accounted dissipating on the lateral four faces of the perturbation. As the fields (2.60) are not dependent on the height, the losses on the top and bottom faces of the perturbation are already accounted in (3.14). Therefore the total power losses for a cavity with a perturbation,  $P_c^p$ , is defined as follows:

$$\begin{aligned} P_c^p = P_c & + \frac{R_s}{2} \left\{ 2 \int_{y=0}^{y=H_{box}} \int_{x=a/2-W/2}^{x=a/2+W/2} |H_x(z=d/2-W/2)|^2 dx dy \right. \\ & \left. + 2 \int_{y=0}^{y=H_{box}} \int_{z=d/2-W/2}^{z=d/2+W/2} |H_z(x=a/2-W/2)|^2 dz dy \right\} \end{aligned}$$

What is more noticeable at this stage, is that the surface resistance,  $R_s$ , accordingly to (3.13), is variable with the frequency at which it is considered: the higher the frequency, the higher the resistance and therefore the higher the conductive losses. As the resonance frequency is made tunable by the insertion of volume perturbations in the cavity, therefore we have to consider the tuned resonance frequency for the evaluation of  $P_c^p$ .

In the case of a cavity with CPP, the computation of the total resistive losses is presented in Figure 3.14. As just mentioned the surface resistance decreases with an increasing CPP volume as the resonance frequency is reduced. This effect is especially visible for small values of the CPP's volume, (i.e.  $W = L < 15$  mm) when the surface of the CPP is relatively small. As the CPP's volume increases, surface losses are proportionally increasing and the reduction of the resistive losses, due to the decrease of the resonance frequency (Figure 3.7) is marginally affecting the losses which are dominated by Ohmic dissipation. For a given value (i.e.  $W = L > 22$  mm) the reduction of the resistive losses, as a consequence of the tuning towards lower frequencies is compensated by the surface on which resistive losses are considered.

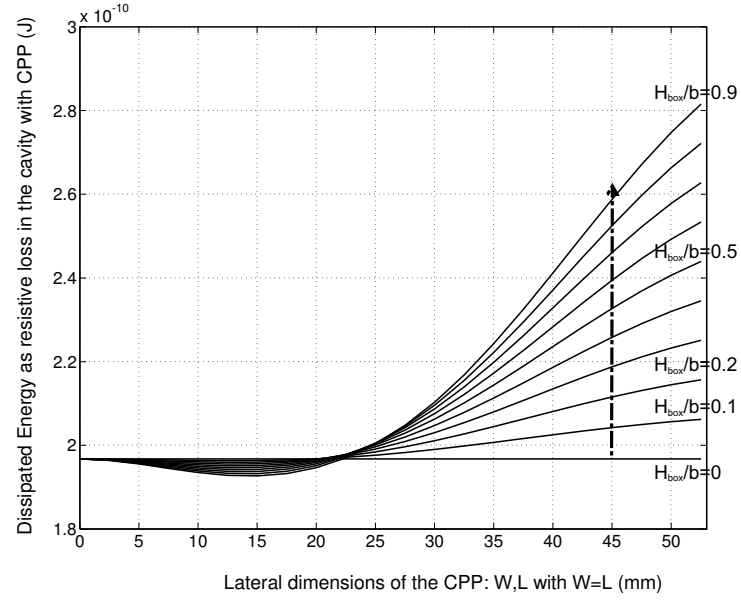


Fig. 3.14.: Total resistive losses in the cavity with CPP.

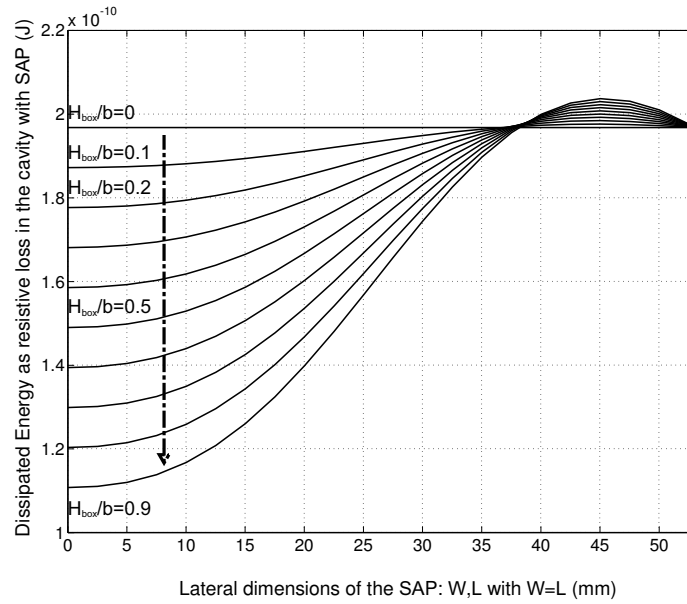


Fig. 3.15.: Total resistive losses in the cavity with SAP.

In the case of a cavity with SAP, the computation of the total resistive losses is presented in Figure 3.15. The first observation concerns the volume of the SAP when the hole located in its center has null dimension (i.e.  $W = L = 0$ ). In this case the resistive losses are a function of the height  $h_{up} = b - H_{box}$  with the SAP volume removed from the cavity volume and losses are computed in the remaining volume. Next, the resistive losses are monotonously increasing with the SAP's volume as the total surface inside the perturbed cavity resonator is increasing. For a given value of the lateral dimensions (i.e.  $W = L > 35$  mm) the effect of the frequency shift due to the insertion of the SAP is made visible as additional losses appear due to the fact that the resonance frequency of the perturbed resonator is higher than in the idle case. Finally, as the side dimensions of the central hole increase further, the resonance frequency of the perturbed resonator is less and less high and the losses decrease, as can be seen in the Figure 3.10.

In the CPP case, the electric energy stored in the perturbed cavity,  $W_{e0_{CPP}}$  is evaluated as follows:

$$W_{e0_{CPP}} = W_{e0} - W_{e_{CPP}} \quad (3.14)$$

where  $W_{e0}$  is the total electric energy stored in the cavity (3.2) and  $W_{e_{CPP}}$  is given by (3.6). The stored electrical energy in the cavity decreases monotonously with an increasing volume of the CPP as can be seen in Figure 3.16. At the maximum lateral dimensions of the CPP the total energy stored corresponds to the energy confined in the volume remaining, with height  $h_{up} = b - H_{box}$ .

In the SAP case, the electric energy stored in the perturbed cavity  $W_{e0_{SAP}}$ , is evaluated as follows:

$$W_{e0_{SAP}} = W_{e_{UP}} + W_{e_{CPP}} \quad (3.15)$$

where  $W_{e_{UP}}$ , (3.9) represents the energy confined in the volume with height  $h_{up} = b - H_{box}$  and lateral dimensions of the cavity.  $W_{e_{CPP}}$  is the total electric energy stored in the CPP (3.6). The electric energy stored in the cavity increases monotonously with increasing lateral dimensions of the hole in the middle of the SAP, as can be seen in Figure 3.16.

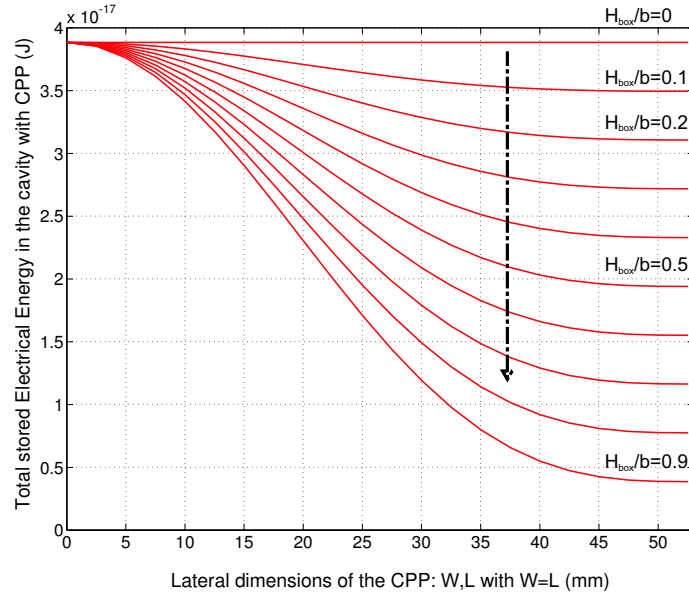


Fig. 3.16.: Total Electrical Energy stored in the cavity with CPP.

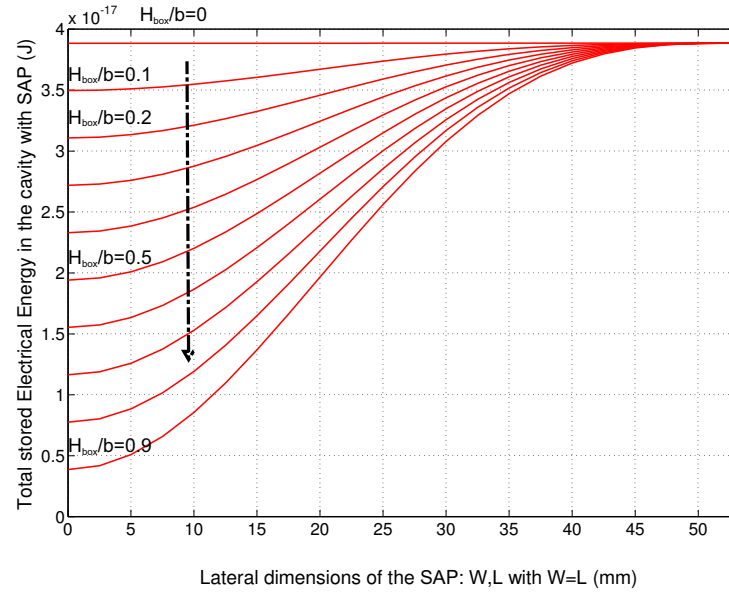


Fig. 3.17.: Total Electrical Energy stored in the cavity with SAP.

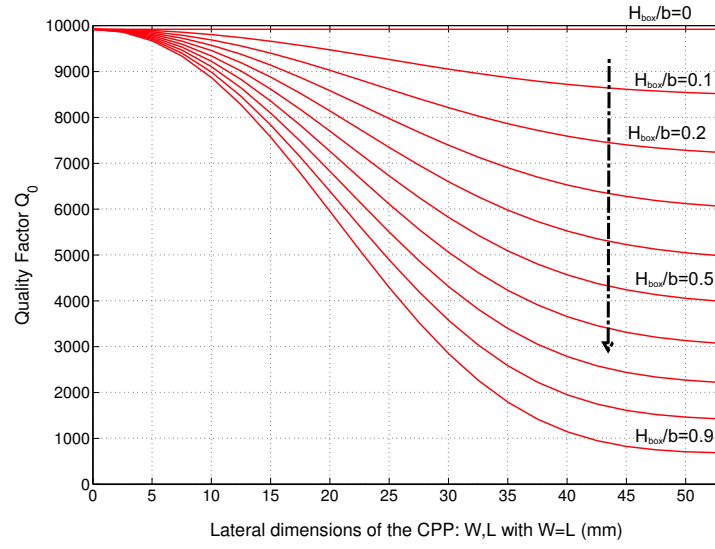


Fig. 3.18.: Quality Factor  $Q_0$  as resulting from the insertion of the CPP.

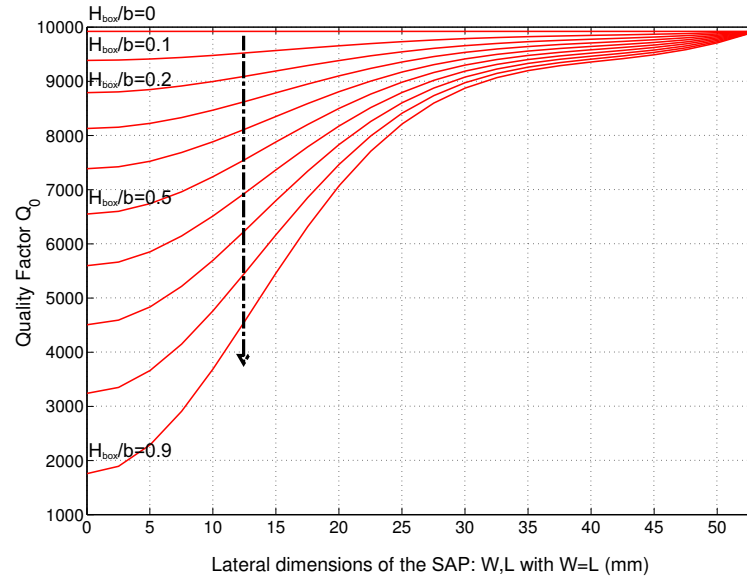


Fig. 3.19.: Quality Factor  $Q_0$  as resulting from the insertion of the SAP.

Finally the Q-factors can be evaluated following (3.11) as the fraction of the energy stored to the resistive losses in the cavity.

In the case of the CPP, the Q-factor decreases from the idle value evaluated in Section 2.4:  $Q_0 = 9946$  for the aluminum walls cavity resonator at 4 GHz. The CPP of a given height,  $H_{box}$  with increasing lateral dimensions  $W = L$  is monotonously reducing the Q-factor. The reduction in the Q-factor is higher for higher values of the CPP's  $H_{box}$ .

In the case of the SAP, increasing values of the lateral dimensions  $W = L$  are increasing the hole in the center of the cavity resonator and therefore reducing the volume of the SAP of a given height,  $H_{box}$ . As a consequence, the Q-factor increases with increasing values  $W = L$  as the dissipated energy is mainly lowering, as can be seen in Figure 3.15 and as the stored energy in the cavity is increasing, as can be seen in Figure 3.17. And it is also obvious that the reduction in the Q-factor is higher for higher values of the SAP's  $H_{box}$ .

### 3.2.5 Q-factor: HFSS simulations and measurements

In this Section we compare the HFSS simulation, measurements and analytical developments presented in Section 3.2.4. In the case of the Q-factor analysis, the behavior and trends are quite similar even though it is hard to plot on the same graph the analytical calculations, HFSS simulations and measurements. First, discrepancies between the measurements and the HFSS simulations were previously mentioned in Section 2.4.2: the measurements provide a much lower unloaded Q-factor than extracted from HFSS simulations, on average  $Q_{0av} = 10350$  vs.  $Q_{0av} = 1969$ , are explained predominantly by additional Ohmic losses encountered at the melding of the metallized PCB with the metallic cavity resonator and due to an excessive contact resistance.

The discrepancy between the analytical evaluation of the Q-factors and HFSS simulation results from the fact that in the analytical evaluation the fields are considered unperturbed as compared to the HFSS simulations, where the field distribution results from the insertion of the perturbation.

In the case of the CPP, the analytical developments are presented in Figure 3.18; the HFSS simulations in Figure 3.20 and the measurements in Figure 3.23. Also, the analytical developments in the case of the SAP, are presented in Figure 3.19; the HFSS simulations in Figure 3.22 and the measurements in Figure 3.23.

In the CPP case, a large difference between simulations and both measurements and HFSS is present for small characteristic dimensions of the post. This is due to the fact that a shorting path to ground appears as a result of the insertion of the post. The effect is particularly high for small lateral dimensions when the idle current distribution is less affected. This effect is not accounted in the analytical evaluation as the idle field distribution is used to compute the losses on the walls of the cavity. The behavior trend is certainly similar as with increasing CPP's lateral dimensions and height the Q-factor is decreasing.



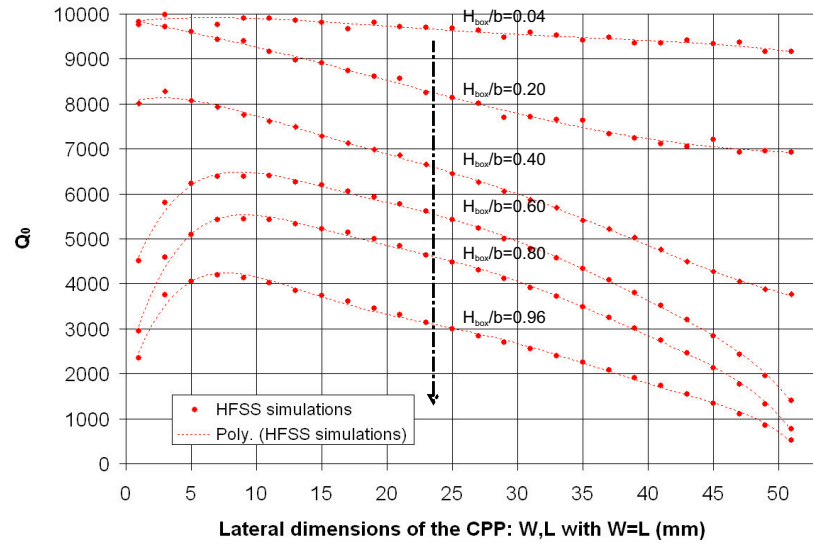


Fig. 3.20.: Quality Factor  $Q_0$  extracted from HFSS simulations corresponding to the insertion of the CPP.

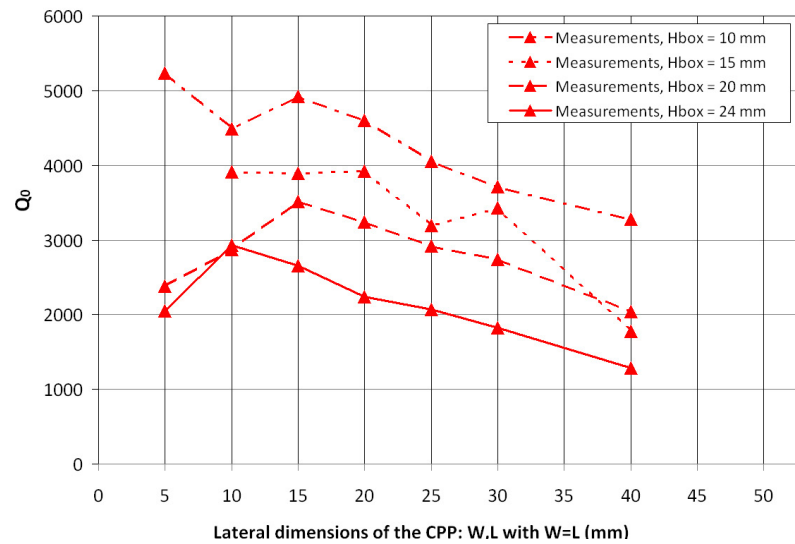


Fig. 3.21.: Quality Factor  $Q_0$  extracted from measurements corresponding to the insertion of the CPP.

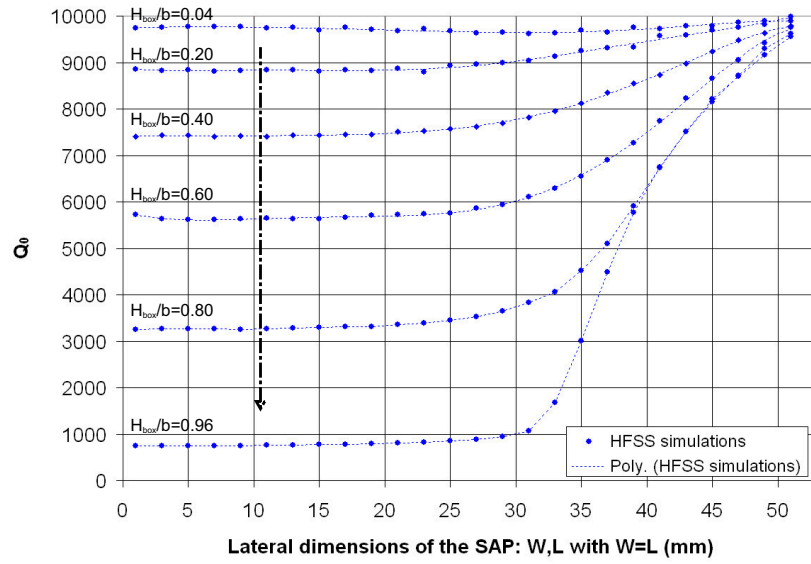


Fig. 3.22.: Quality Factor  $Q_0$  extracted from HFSS simulations corresponding to the insertion of the SAP.

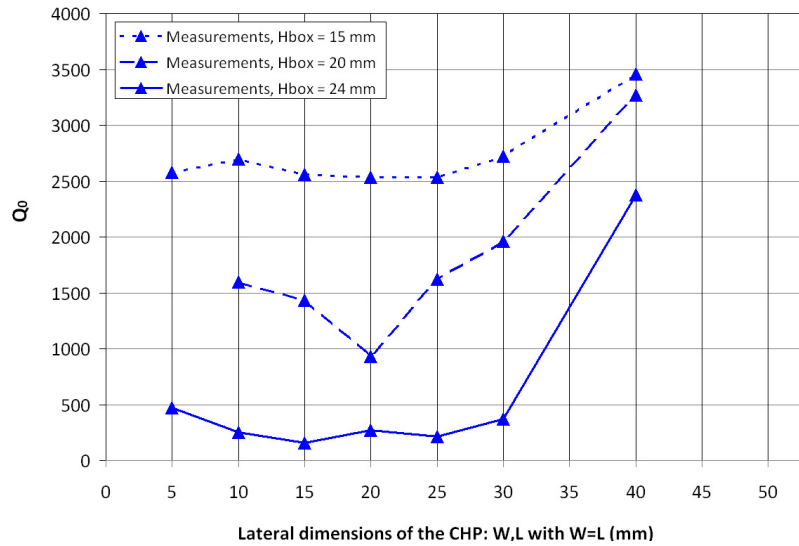


Fig. 3.23.: Quality Factor  $Q_0$  extracted from measurements corresponding to the insertion of the SAP.

In the SAP case, the impact of the internal volume of perturbation is very similar in HFSS simulations and measurements. The Q-factor is monotonously increasing with increasing SAP's hole lateral dimensions and is decreasing with an increasing volume's height. The main difference with simulations is that the Q-factor is increasing faster considering the lateral dimensions of the SAP's hole, i.e. starting from ( $W = L = 5$  mm) to ( $W = L = 30$  mm). This is a consequence of the fact that the electric field does not engulf in the hole of the SAP's volume, as the dimensions of the hole are very small in comparison with the wavelength. This effect is clearly visible in Figure 3.11-c). When dimensions  $W, L$  are large enough the field engulfs in the hole and a higher stored electrical energy is made possible for the cavity which results in a higher Q-factor. In the analytical evaluation this effect is not accounted as the energy is evaluated considering the fields of the idle case.

### 3.3 IMPLEMENTATION AT 60 GHZ: INTEGRATION IN MCM-D

This Section is summarizing the research performed within the European Project 3D $\mu$ Tune. This project aimed to industrialize a tunable integrated cavity resonator characterized by a high Q-factor in the range of 500-1000, with a resonance frequency in the range 10-110 GHz and provided with a tunable resonance frequency up to 5% i.e. from 59 to 62 GHz. Within the benchmark of RF resonators this component is filling a free spot, as presented in Figure 3.24. Different available devices are presented for comparison regarding their resonance frequency and Q-factor. Those devices are the T-lines: Transmission line resonators, BAW: Bulk Acoustic Wave resonators, SAW: Surface Acoustic Wave resonators and DR: dielectric resonators. Nanotimer is a mechanical silicon resonator developed at IMEC.

More precisely, the project focused on three different demonstrators in the millimeter-wave range: RADAR-based Adaptive Cruise Control (ACC) next-generation system at 74.3 GHz, Satellite Communication with a 20 GHz passband filter for satellite transceivers and 60 GHz PLL-free VCO for future local area networks (WLAN).

Within the 3D $\mu$ Tune project, a generic production process towards reliable, fixed high Q-factor resonators was developed. Microfabrication technologies are used to realize a cavity resonator which is etched in silicon, metallized and finally implemented above the existent MCM-D platform. The research undertaken was clearly focusing on the development of an integrated means towards the realization of the tunability. The interest in developing tunable cavity resonators resides in the fact that one single component could be used instead of many others as it is actually the case. Therefore, systematic and non systematic errors about the frequency resolution could be solved in the operation time thanks to the tuning capability. Furthermore, expensive manual handling could also be avoided using the trimming capability that is offered. Finally, those resonators are considered to be combined with BiCMOS Silicon Germanium (SiGE) MMICs in a hybrid multi-technology module towards the realization of tunable cavity resonators based oscillators.

Within the framework of my PhD, I focused on the development of 60 GHz tunable cavity resonators implemented in IMEC's MCM-D platform, results of which are presented in this Chapter.

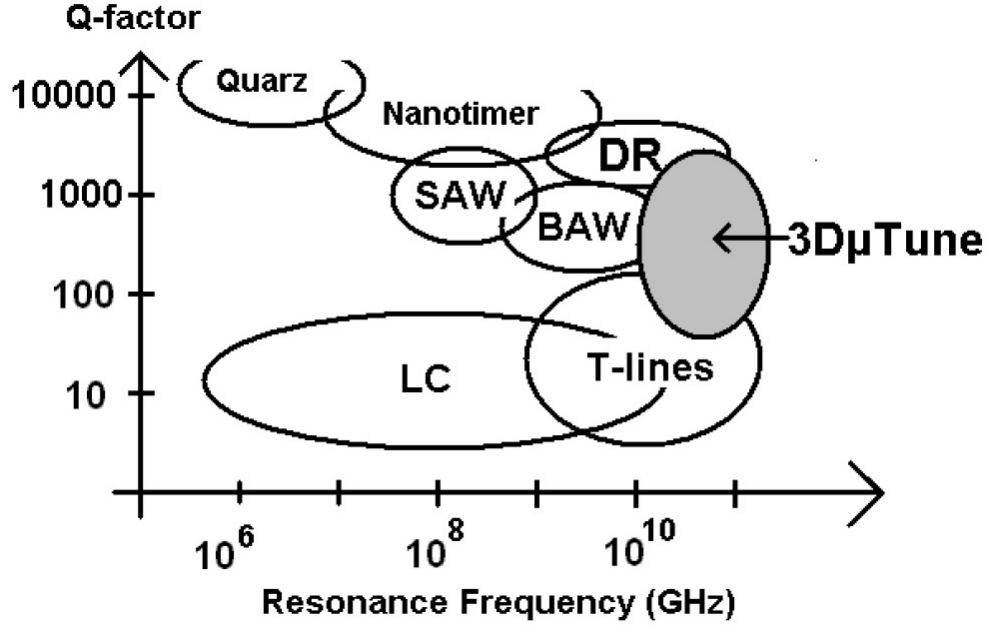


Fig. 3.24.: Q-factor versus resonance frequency for a variety of resonator technologies in comparison with the European Project 3D $\mu$ Tune. (T-lines: Transmission line resonators, BAW: bulk acoustic wave resonators, SAW: surface acoustic wave resonators, DR: dielectric resonators, Nanotimer: mechanical silicon resonators developed in the EU project Nanotimer). Source 3D $\mu$ Tune project (IST-2005-027768) and IMEC.

### 3.3.1 Principle of design and modelling

This section reviews the design and modelling of a 60 GHz tunable cavity resonator. The controllable volumes located inside the cavity are shaping a Faraday cage formed by the upward curvature of electrostatic MEMS inspired by the EFFA actuators developed by [Rottenberg 07]. The small perturbation theory described in Section 3.2.2 is used to study the resonance behavior of the perturbed cavity resonator. The resonance frequency is computed following the integral method described in Section 3.2.2. The results of the analytical analysis concerning a cavity resonator with lateral dimensions  $3500 \times 3500 \mu\text{m}$  and a height of  $440 \mu\text{m}$ , corresponding to a resonance of 60 GHz, are presented in Figure 3.25. The effect of both CPP and SAP on the resonance frequency is described and compared to the 3D $\mu$ Tune project's specifications, i.e. 59-62 GHz and the available unlicensed frequency band around 60 GHz, i.e. 57-64 GHz. The parameters analyzed are sweeping the lateral dimensions of the CPP, SAP from 0 to  $3500 \mu\text{m}$  and the height of the perturbation from 0 to  $120 \mu\text{m}$ , by steps of  $10 \mu\text{m}$ .

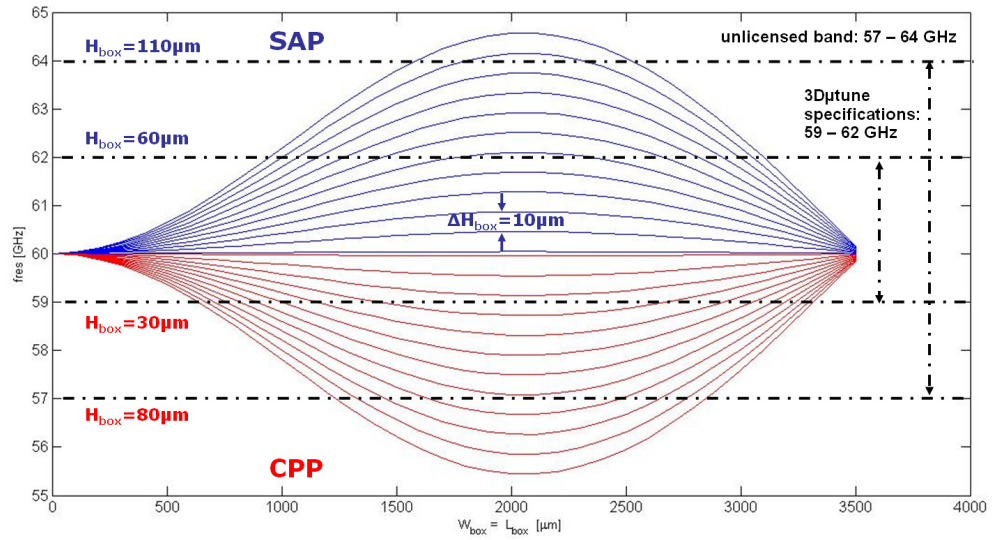


Fig. 3.25.: Comparison with 3D $\mu$ Tune project's specifications and the available unlicensed band around 60 GHz. The height of the perturbation is swept from 0 to 120  $\mu m$  by steps of 10  $\mu m$ .

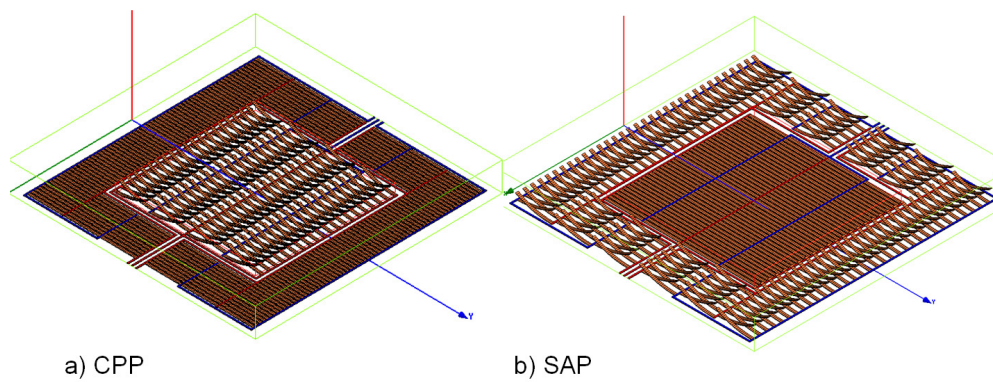


Fig. 3.26.: Concept of dual actuation of both CPP and SAP by EFFA MEMS located inside the cavity resonator.

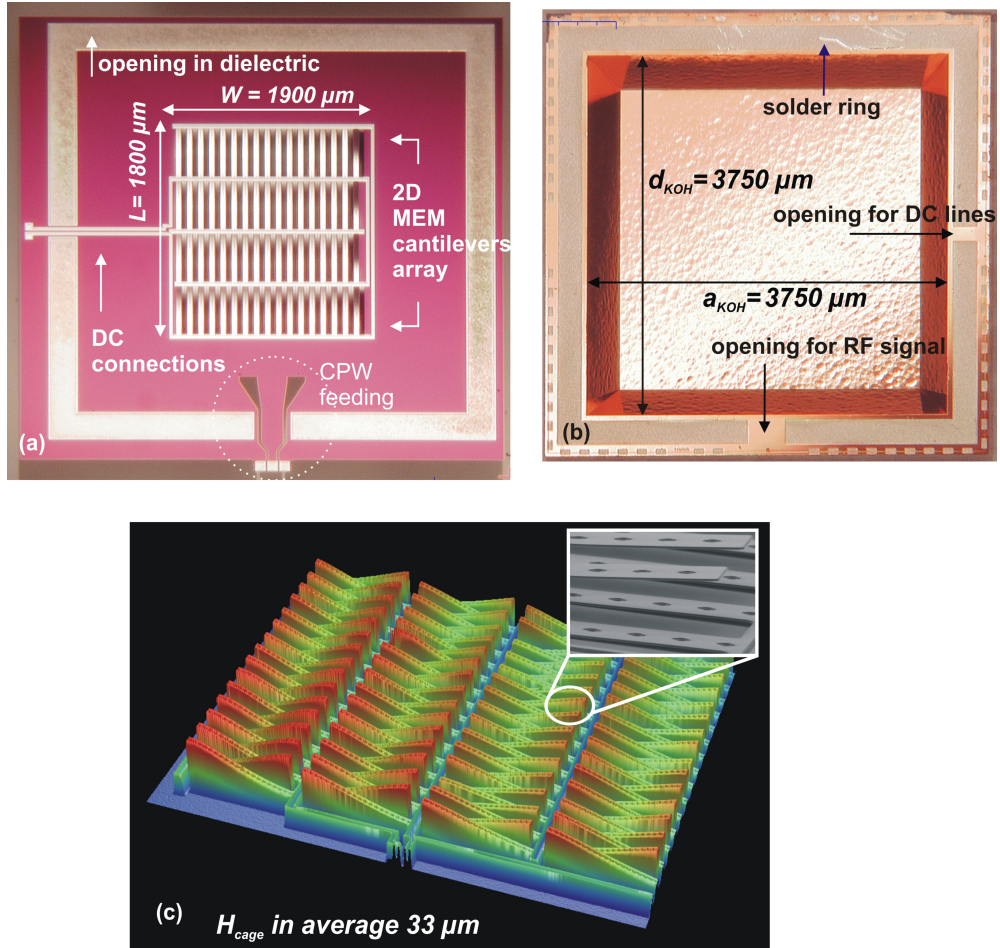


Fig. 3.27.: Photograph of the (two parts of the) fabricated MEMS tunable cavity resonator: a) bottom part comprising the CPW feeding and the array of cantilevers located at the center; b) top part with the KOH-etched and metallized cavity. c) Optical profilometer plot and a SEM close-up of the cantilevers array implementing the Faraday cage, located inside the cavity resonator. The colors represent the out-of-plane deflection. The ( $500 \mu\text{m}$  long) cantilevers display a tip deflection, on average of  $33 \mu\text{m}$ .

The Faraday cage implementation with MEMS is based on the fact that a volume surrounded by conductive walls is defining an equipotential enclosure, a Faraday cage. Due to the external electric fields, forces are produced on the charges within the conductor walls, forcing the electrons to rearrange accordingly. The new charge distribution generates an electric field inside the cage which neutralizes the external field. Even in case of openings in the Faraday cage wall, an effective shield against external



electromagnetic fields is formed provided that the openings are much smaller than the wavelength (so that there is no coupling through the openings).

In our implementation, the Faraday cage is formed by an array of cantilevers curled upwards, schematically shown in Figure 3.28 and provided with small openings in between two cantilevers ( $5 \mu m$ ). The cantilever form the EFFAs and are characterized by the following dimensions: width =  $50 \mu m$ , length =  $500 \mu m$  and slot =  $5 \mu m$ , referring to [Rottenberg 07].

Differently, the cantilevers are isolated from the bottom electrode by a high permittivity dielectric layer (200 nm AlN 200 nm, relative permittivity  $\epsilon_r = 9$ ). This implementation preserves the RF integrity of the resonator while ensuring a DC actuation of the cantilevers array. The deflection of the cantilevers defines the height  $H_{cage}$  of the Faraday cage and can be controlled by applying a DC voltage up to 20 V, between the cantilevers and the ground or between the two sets of fingers.

Figure 3.27 shows a photograph of the fabricated bottom part a) and top part b) of the cavity resonator and an optical profilometer plot and SEM photograph c) of part of an actually fabricated array located in the middle of the cavity resonator. The effective shielding of the array of cantilevers is equivalent to a rectangular volume with lateral dimensions equal to the array lateral dimensions and with a height equal to the tip deflection of the cantilevers (as depicted in Figure 3.28).

The volume defined by the MEMS is variable and thus the electric or magnetic stored energy is perturbed, which results in a shift of the resonance frequency of the cavity resonator. By placing the volume in the center of the cavity, the resonance frequency is decreased, whereas placing the volume at the circumference, the resonance frequency is increased, configurations were called respectively CPP and SAP in Section 3.2.1.

As a direct consequence of the project specifications and for the sake of the research, both CPP and SAP zones were considered for actuation. Actually this choice was also motivated by the limitations in the excursion of the height of the available MEMS (max.  $50 \mu m$ ), thus limiting the tuning by a single kind of array and doubling it, by using both kinds of arrays.

A schematic view of the proposed concept of tunable cavity resonator, implementing a CPP is depicted in Figure 3.28. Shown is a rectangular cavity (with dimensions  $a \times d \times b$ ), in which a rectangular Faraday cage (with dimensions  $W \times L \times H_{cage}$ ) is placed at the bottom-center. The cavity is wet-etched in silicon with sloped and metallized lateral walls tilted by an angle of  $54.75^\circ$ .

As previously introduced in Section 2.4.1, among the rectangular cavities, a square cavity (i.e.  $a = d$ ) provides the highest possible unloaded Q-factor,  $Q_0$ . The Q-factor scales linearly with the height  $b$ .

$$a = \frac{c_0}{\sqrt{2}f_{101}} \quad (3.16)$$

with  $c_0$  the speed of light.

However, in the fundamental mode  $TE_{101}$ , the height  $b$  is not influencing the resonance frequency. In the case of a KOH-etched cavity, with slanted side walls, the resonance frequency of the  $TE_{101}$  is well influenced by the height. The new resonance frequency

is approximated by the mean values between the bottom (smaller) and top (larger) lateral dimensions of the slanted cavity. The trapezoidal cavity will have larger dimensions than a square one, in order to resonate at the same frequency. The general equation (3.16) was therefore adapted to the trapezoid KOH-etched cavity [Ocket 08]:

$$a_{KOH} = a + \frac{b}{\tan(54.75^\circ)} \quad (3.17)$$

With a targeted resonance frequency  $f_{101} = 60$  GHz the lateral dimensions of the KOH-etched cavity are  $a_{KOH} = d_{KOH} = 3844 \mu m$ , for a given height  $b = 440 \mu m$ . The maximal height of the cavity is set by the maximal depth before making the wafer brittle and is a consequence of the fabrication constraints. The CPP is formed by a Faraday cage composed of cantilevers which perturb the electromagnetic field distribution, thereby affecting the resonance frequency of the cavity resonator in its fundamental  $TE_{101}$  mode of resonance. The size of the Faraday cage determines the shift in frequency. By making the cage height  $H_{cage}$  variable, the resonance frequency can be tuned. Perturbations are centrally located for obvious symmetry reasons considering the field distribution of the electrical field in the  $TE_{101}$  mode.

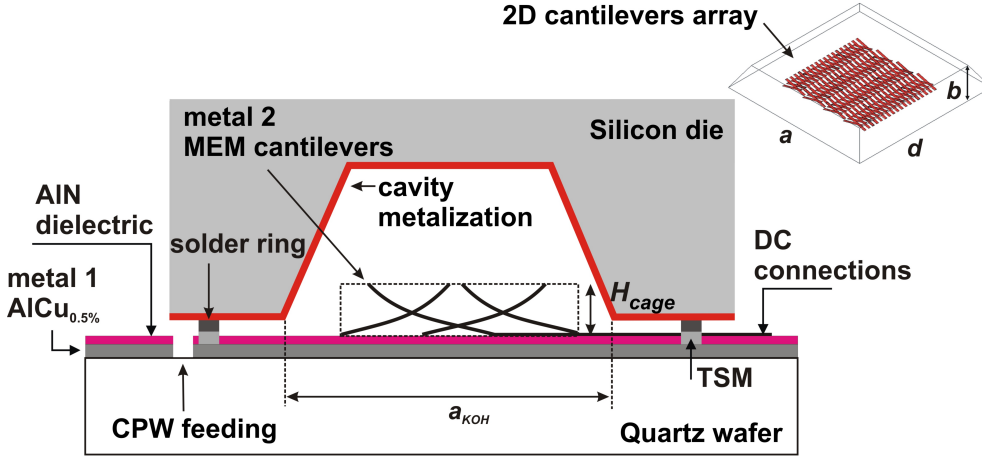


Fig. 3.28.: Cross-section of the MEMS tunable cavity resonator. The resonator is formed by a two-chip assembly: the bottom part comprises the CPW feeding and the top part is composed by the KOH-etched and metallized trapezoid cavity. Inset of the schematic 3D structure.

The frequency shift produced by a perturbation located inside the cavity is analytically evaluated following the small perturbation theory [Pozar 05], described in Section 3.2.2 for both CPP and SAP. The computation of the frequency shift is based on the evaluation of the electrical  $W_e$  and magnetic  $W_m$  stored energy in both the perturbed and initial volume of the cavity. An approximation of the frequency shift is given by the fraction of the stored energy in the perturbed volume related to the stored energy in



the initial non perturbed (3.4). The frequency shift provided by a rectangular metallic box or equivalently by the Faraday cage with lateral dimensions  $W = L = 1900 \mu m$ , located in the center of the square rectangular cavity resonator (as in Figure 3.28) with nominal dimensions  $a_{KOH} = d_{KOH} = 3844 \mu m$  and  $b = 440 \mu m$  is calculated in Figure 3.29. It is shown that the larger the height of the metallic box,  $H_{cage}$ , the larger the shift in frequency.

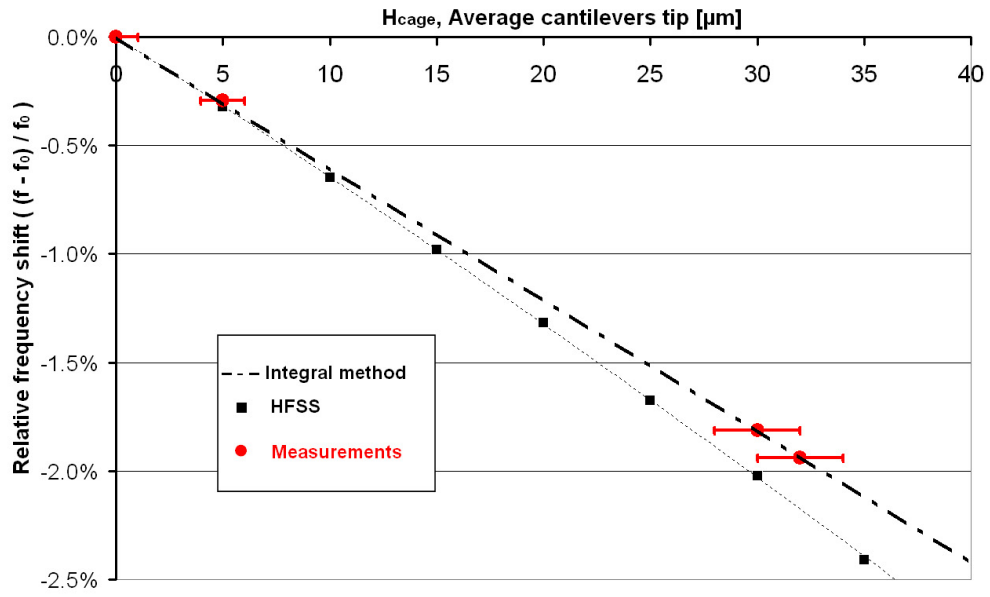


Fig. 3.29.: Comparison between the analytical method (3.4), HFSS simulations and measurements of the frequency shift as a result of the cavity perturbation by a CPP. The simulations apply for a rectangular cavity resonator with dimensions:  $a = d = 3844 \mu m$  and  $b = 440 \mu m$ , and inner Faraday cage dimensions  $W = L = 1900 \mu m$ , with variation of the height  $H_{cage}$  from 0 to  $40 \mu m$ .

The integral method used to evaluate the frequency shift as a result of the rectangular perturbation, differs by only 0.5% from an HFSS simulation for a cage with a height of  $40 \mu m$  in a cavity resonator with a height of  $440 \mu m$ . The deviation is a result of the hypothesis used in the analytical evaluation for which the fields are considered unperturbed, something which is not the case in the HFSS simulations, where the fields are computed for a given perturbation. The measurements are also included in Figure 3.29, and are performed on a trapezoid cavity resonator demonstrator (see Figure 3.28 and Figure 3.27) and discussion in Section 3.3.6.

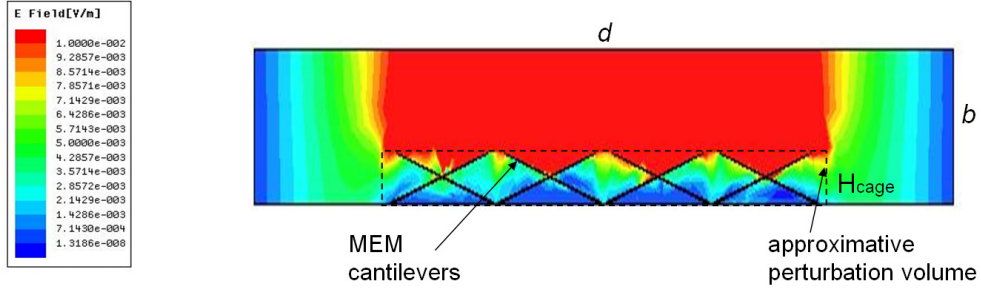


Fig. 3.30.: E field distribution in a cross section inside a cavity resonator provided with EFFA MEMS perturbation volume. The approximation volume ( $W \times L \times H_{cage}$ ) is also indicated.

As can be seen in the cross-section of Figure 3.28, the Faraday cage is defined by a 2D array of electrostatically controllable metal cantilevers. The cage perturbs the electromagnetic field distribution inside the cavity resonator and provides a means to electrically tune the resonance frequency. The effective shielding of the MEMS Faraday cage is somehow lower than the height of an equivalent metallic box, with the height equal to the tip height of the cantilevers, as can be seen in Figure 3.30.

The HFSS simulation curve lies below the integral method, meaning that the frequency shift following the evaluation by the HFSS is larger than the one evaluate analytically. The reason was previously detailed in Section 3.2.3, as resulting from the non homogeneity of the E field, in reality and considered so by HFSS. If in Figure 3.29 the measurements are getting closer to the analytical evaluation, this is not because the integral method is more accurate but because the effective shielding of the MEMS array is smaller than the one provided by a full metallic box, as can be seen from the field distribution presented in Figure 3.30.

At higher frequencies, the higher the resonance frequency, the higher the frequency shift for a MEMS actuator characterized by a given deflection height. The reason is that with higher frequencies, the size of the cavity resonator is reduced and the impact of the MEMS deflection is increased. Actually, the lateral dimensions of a cavity resonator are  $a = d = 770 \mu m$  at 300 GHz;  $460 \mu m$  at 500 GHz and  $230 \mu m$  at 1 THz. The opposite is valid for lower frequencies.

### 3.3.2 Fabrication process

The fabrication process involves two separate parts: the first part, see Figure 3.27-a), is a carrier substrate comprising the RF feeding, the MEMS array and the DC feeding for the actuation of the electrostatic MEMS. The second part, see 3.27-b) is a die, in which a cavity is etched and metallized and which is solder bonded onto the carrier wafer to form the cavity resonator.

### 3.3.2.1 Fabrication of the carrier substrate

Designs were finalized in layouts for an initial process flow, illustrated in Figure 3.31. Visible is the process flow of the bottom part composing the cavity resonator.

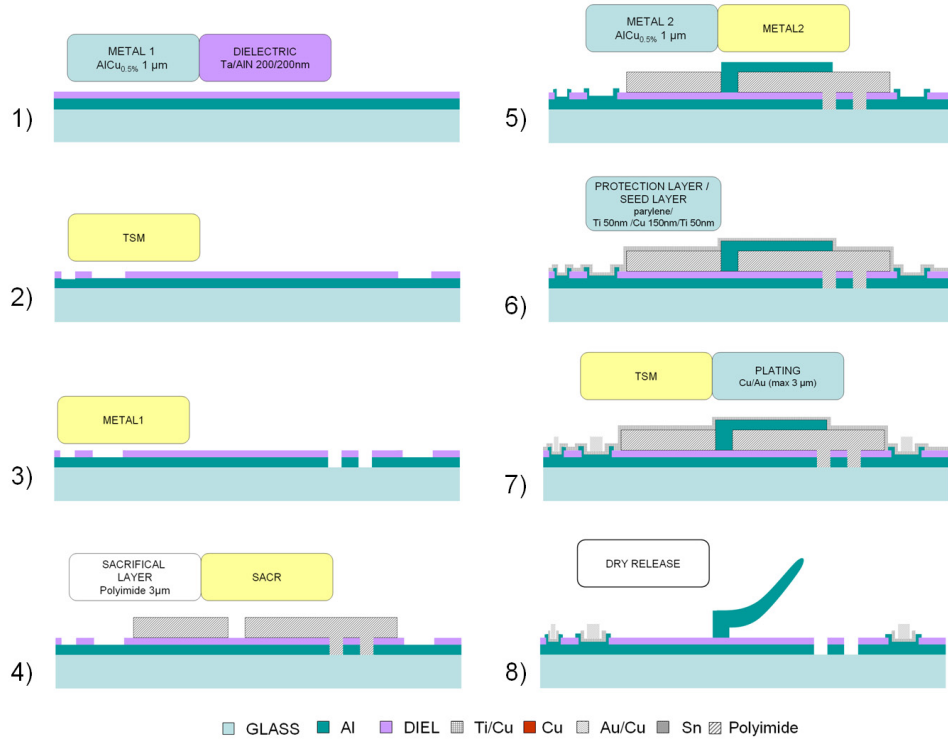


Fig. 3.31.: Process flow description (steps 1 to 8) of the bottom structure. Source 3D $\mu$ Tune project (IST-2005-027768) and IMEC.

To realize the cavity resonator two main processes are involved using in total eight masks: three masks are needed for the cavities and five to realize the landing structure, with MEMS. The fabrication of the bottom part starts with a quartz carrier wafer. 1) A first metallization of 1  $\mu\text{m}$  copper aluminum alloy,  $\text{AlCu}_{0.5\%}$  is deposited and is covered by a layer of 200 nm of tantalum oxide  $\text{Ta}_2\text{O}_5$  or  $\text{AlN}$ . The oxide layer allows grounding the MEMS array for RF while insulating for DC. The successive steps in the processing are 2) the patterning of the dielectric followed by 3) the patterning of the first metallization to shape the RF feeding. 4) Then a sacrificial layer of 2.5  $\mu\text{m}$  polyimide film is used as sacrificial layer for 5) the 2<sup>nd</sup> metallization, a 1  $\mu\text{m}$  thick  $\text{AlCu}_{0.5\%}$  layer which is patterned to form the cantilever structures. 6) Next, a Ti 50 nm / Cu 150 nm / Ti 50 nm seed layer is covering the whole surface in order to deposit the Top Surface Metallization (TSM) of Cu 3  $\mu\text{m}$  / Au 150 nm 7) by electroplating. Above the seed layer a parylene layer is deposited and patterned to protect the cantilevers during the TSM electroplating step. 8) After the seed layer

removal, an  $O_2$  plasma etching is used to etch the polyimide, thereby releasing the MEMS. The ring of the TSM with a Au finish (i.e., the metallization to which the soldering is done) is interrupted on two locations: one to allow for the RF coupling and the other for the connectivity to the MEMS actuator. The process flow was developed in a succession of trials and errors among which representative issues are presented in Figure 3.32.

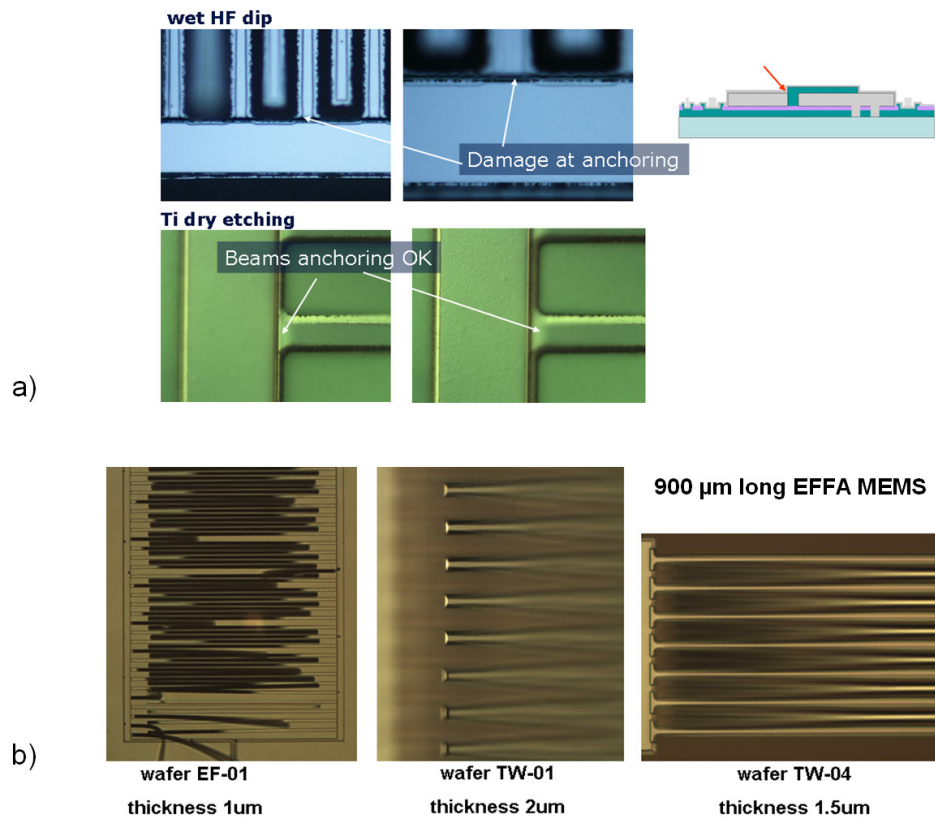


Fig. 3.32.: a) Damages at the anchoring of the MEMS due to the HF dip to remove the seed layer and b) thickness and length variation of the cantilevers.

The first issue, was the stiction of the cantilevers because of extra charges trapped in the dielectric layer underneath the beams. Using aluminum-nitride AlN instead of tantalum oxide  $Ta_2O_5$ , as a insulation dielectric, fixed the problem. The second issue was the compatibility between the actuator integrity and the TSM process. Actually, starting from a basic two metallization MEMS process as required for the realization of the EFFA, following [Rottenberg 07], a TSM solder ring had to be added for packaging. The plating of this ring involves the deposition of a seed layer and the removal of the seed layer, using a wet HF dip, was harmful for the anchoring of the MEMS (Figure 3.32-a)). The solution found was to use a dry etch to remove the seed

layer. The third issue was solved by providing openings in the dielectric layer, in order to be able to have appropriate RF measurements as the dielectric was covering the first metallization at the probing pads. The dielectric layer is also etched at the locations of the connection between the cavity and the bottom metallization. The fourth issue occurred during the dicing of the cavity wafers as the copper layer delaminated. A new process step was introduced in order to pattern the copper. Finally, the  $900\text{ }\mu\text{m}$  long cantilevers stated to be too thin ( $1\text{ }\mu\text{m}$ ) to maintain their rigidity or too thick ( $1.5, 2\text{ }\mu\text{m}$ ) and impossible to actuate because of their stiffness, in Figure 3.32-b). The next developments involved shorter beams,  $500\text{-}400\text{ }\mu\text{m}$  long cantilevers known to actuate and are presented in Section 3.3.3.2.

### 3.3.2.2 Fabrication of the cavities substrate

The process flow concerning the realization of the cavities is described in Figure 3.33. The top part processing, shaping the cavity starts with a  $675\text{ }\mu\text{m} < 100 >$  low resistivity silicon wafer, onto which  $200\text{ nm}$  LPCVD SiN is deposited and into windows are dry etched. These openings serve as a hard mask for the KOH etching of the cavities. After the  $440\text{-}460\text{ }\mu\text{m}$  deep cavities are etched they are covered by  $1\text{ }\mu\text{m}$  thick sputtered Cu layer. A Cu/Sn ( $5\text{ }\mu\text{m} / 9\text{ }\mu\text{m}$ ) solder ring is locally electroplated. This solder ring matches the connection ring on the bottom carrier TSM side. The cavity wafer is diced and the individual dies are flip-chip assembled onto the carrier using a low temperature soldering process.

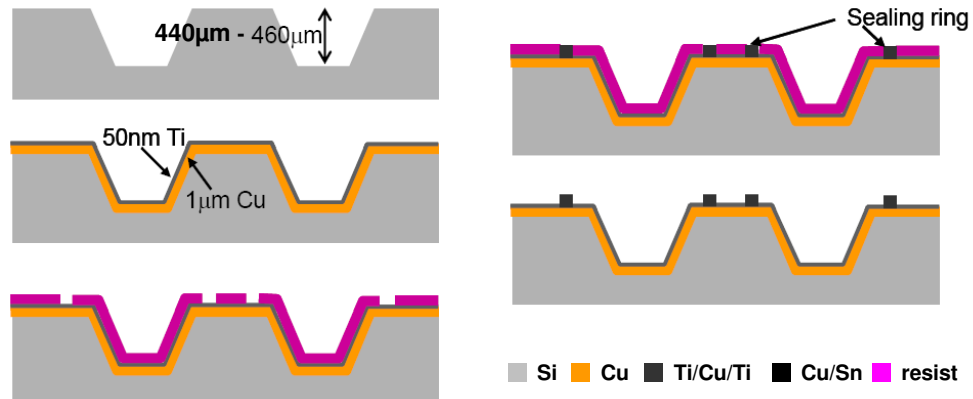


Fig. 3.33.: Cavity wafer process flow. Source 3D $\mu$ Tune project (IST-2005-027768) and IMEC.

### 3.3.3 Design overview

The implementations were realized in two technologies. The first technology was used to demonstrate the tunability of cavity resonators by electrically actuated MEMS and is composed of two iterations: the DEMON and the FAKIR run. The second technology, the MCM-D is the technology in which such devices are aimed to be finally

integrated. The development of high Q-factor cavity resonator at 60 and 75 GHz was therefore realized to demonstrate the integration of air cavity resonators and is called the MONOCEROS run. The coplanar feeds developed in Sections 2.4.3 and 2.4.4 at 4 GHz, were adapted to the technologies above mentioned.

### 3.3.3.1 First demonstrator: DEMON run

A first iteration run, called DEMON, see Figure 3.34 was designed in order to prove that a cavity resonator can be made tunable by using an array of MEMS located inside the resonator and actuated by electrical means.

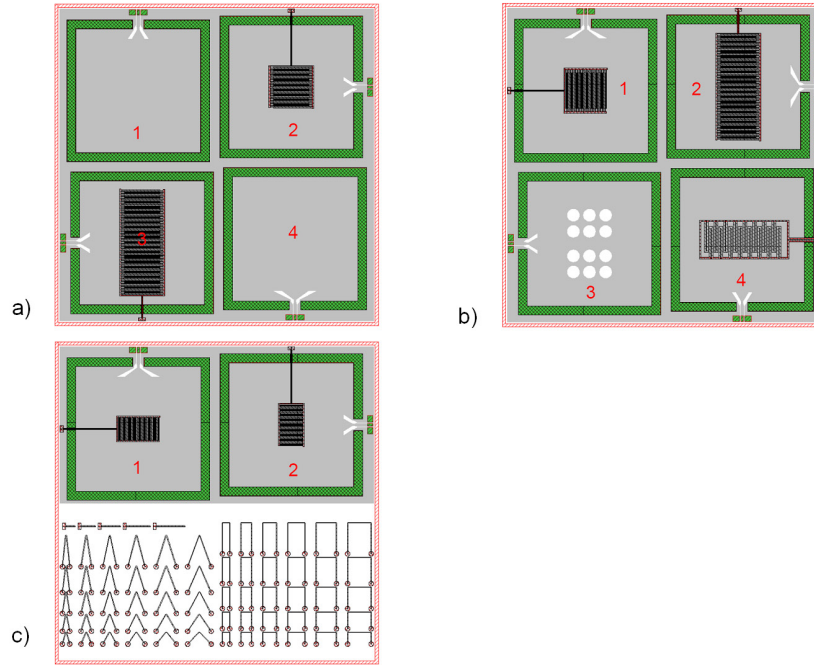


Fig. 3.34.: Three cells designed for the DEMON run, incorporating ten different cavity resonators with and without MEMS tuning device.

This run was realized while further investigations concerning the development of the theory were undergoing, therefore the perturbation volume was not optimized at that time and is declined in two versions of CPP. The first CPP has lateral dimensions  $W = L = 1300 \mu m$  and is presented in Figure 3.34 at locations a)-2, b)-1 and c)-1-2. HFSS Simulations of the cavity resonator, tuned by a metallic box representing the CPP version 1, considering the height of perturbation metallic volume variable from 1 to  $31 \mu m$ , results in a maximal frequency shift of 1 GHz for  $H_{cage} = 31 \mu m$ . The second CPP has lateral dimensions  $W = 1300 \mu m$ ,  $L = 3300 \mu m$ , and is presented in Figure 3.34 at locations a)-3, b)-2 and c)-1. Simulations of the cavity resonator with this CPP, version 2, result in a typical frequency shift of 0.6 GHz for  $H_{cage} = 31 \mu m$ .

This is related to Section 3.2.2, where a CPP with a width and length close to half the cavity's lateral dimensions, is providing the maximum frequency shift.

Two different feeds were designed for the run as presented in Figure 3.35 where the characteristic dimensions are provided. The first feed is designed to be slightly over-coupled and is characterized by Q-factors of:  $Q_0 = 803$ ,  $Q_e = 259$ ,  $Q_l = 196$  and  $\kappa = 3$ . The Q-factors are extracted from HFSS simulations using the QZERO software. The layouts with this feeding are presented in Figure 3.34, at the locations a)-1-2-3, b)-3 and c)-2. The second feeding is designed to be largely overcoupled and is characterized by Q- factors of:  $Q_0 = 505$ ,  $Q_e = 37$ ,  $Q_l = 40$  and  $\kappa = 12$ . The layouts with this feed are presented in Figure 3.34, at the locations a)-4, b)-1-2 and c)-1.

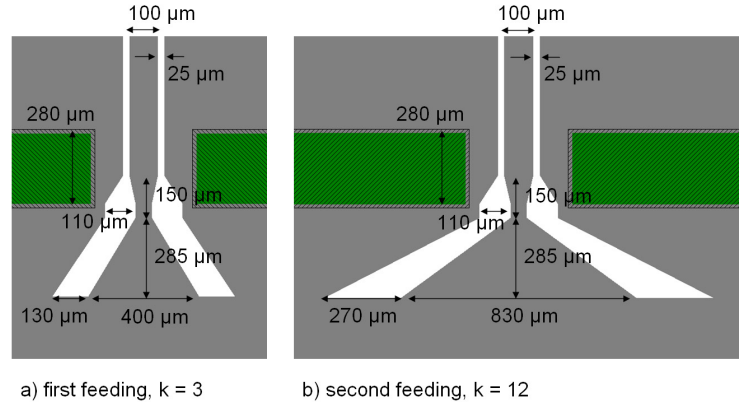


Fig. 3.35.: Layouts and characteristic dimensions of the feeds designed for the DEMON run. Adaptation of Sections 2.4.3 and 2.4.4.

### 3.3.3.2 Optimized demonstrator: FAKIR run

While designing the FAKIR run, the process flow was adapted. The new flow/design incorporates 400-500  $\mu\text{m}$  long cantilevers with a width of 20  $\mu\text{m}$  and a spacing of 5  $\mu\text{m}$  (between the cantilever fingers). The new run was set to demonstrate the feasibility of the tuning by an array of MEMS forming a Faraday cage inside the cavity resonator as this was not yet demonstrated. Concomitantly, the actuation behavior of the EFFA MEMS was further investigated by including in the run a large variation of length, width and spacing characterizing the cantilevers. A parametric study of 210 variations were designed as follows: length (400, 500, 600, 700, 900, 1100, 1200  $\mu\text{m}$ ), width (20, 25, 30, 40, 50  $\mu\text{m}$ ) and slot (2, 3, 4, 5, 7, 9  $\mu\text{m}$ ). Six new feeds were designed and their characteristics are presented in Table 3.1. The design of these feeds is based on the developments presented in Section 2.4.4: the longer the CPW inside the cavity resonator, the higher the coupling coefficient and the higher the coupling the larger the aperture around the shorted CPW line. Finally, the couplings are optimized with HFSS, considering the slanted walls KOH-etched cavity resonator. The couplings are



of under-coupled, critically coupled and over-coupled. The layouts are presented in Figure 3.37.

Table 3.1.: Designs of the FAKIR run. Coupling coefficient,  $\kappa$ , unloaded Q-factor,  $Q_0$  and loaded Q-factor,  $Q_l$ .

| feed | $\kappa$ | $Q_0$ | $Q_l$ |
|------|----------|-------|-------|
| 1    | 0.18     | 1055  | 898   |
| 2    | 0.36     | 1056  | 776   |
| 3    | 0.95     | 922   | 473   |
| 4    | 1.84     | 994   | 350   |
| 5    | 3.9      | 1073  | 218   |
| 6    | 9.8      | 565   | 53    |

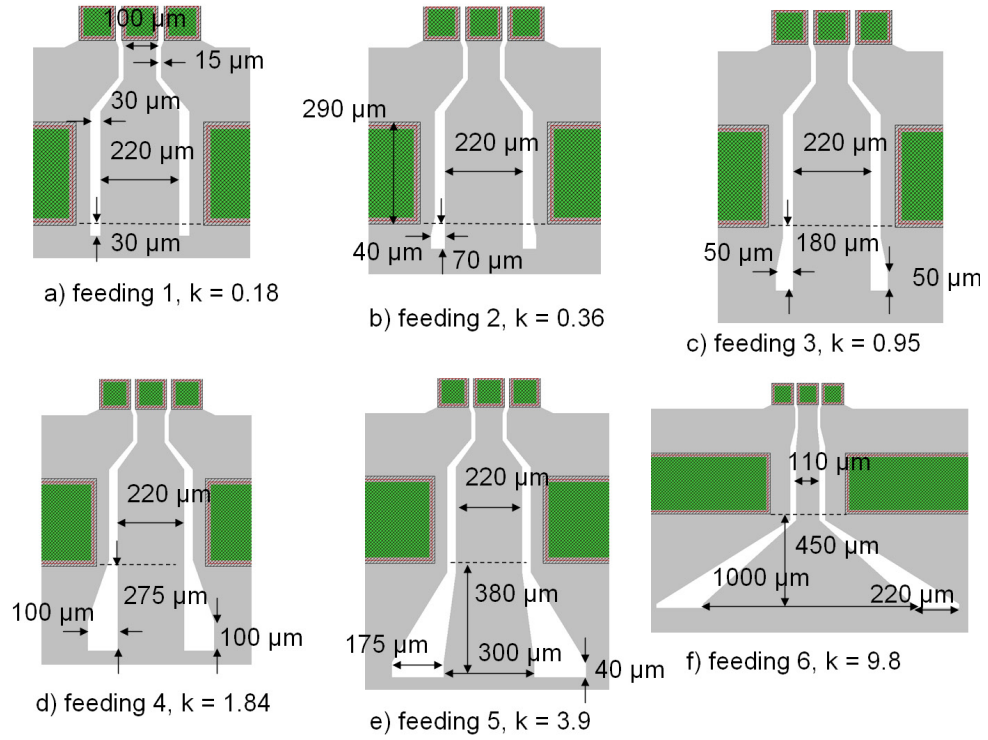


Fig. 3.36.: Layouts and dimensions of the six feeds of the FAKIR run.



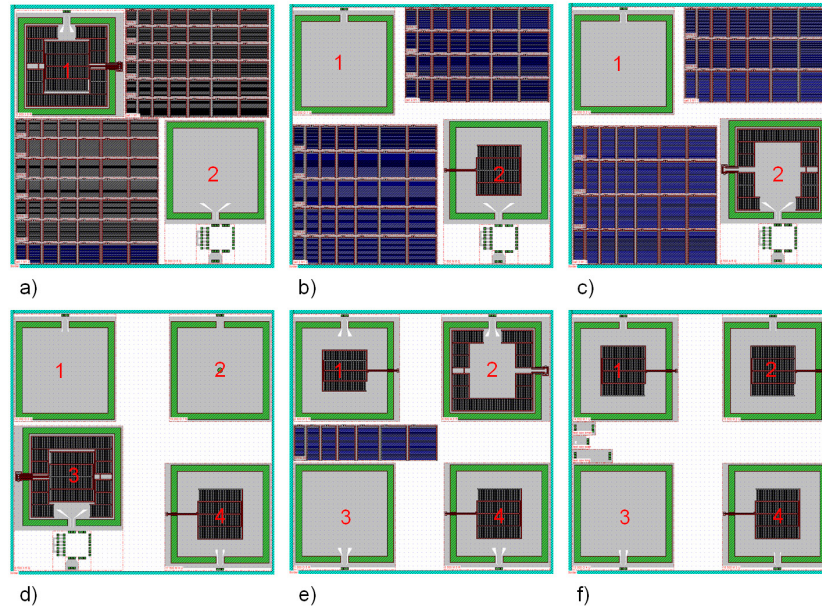


Fig. 3.37.: Layouts of the six cells of the FAKIR run (a-f) incorporating eighteen different cavity resonators and volume perturbations.

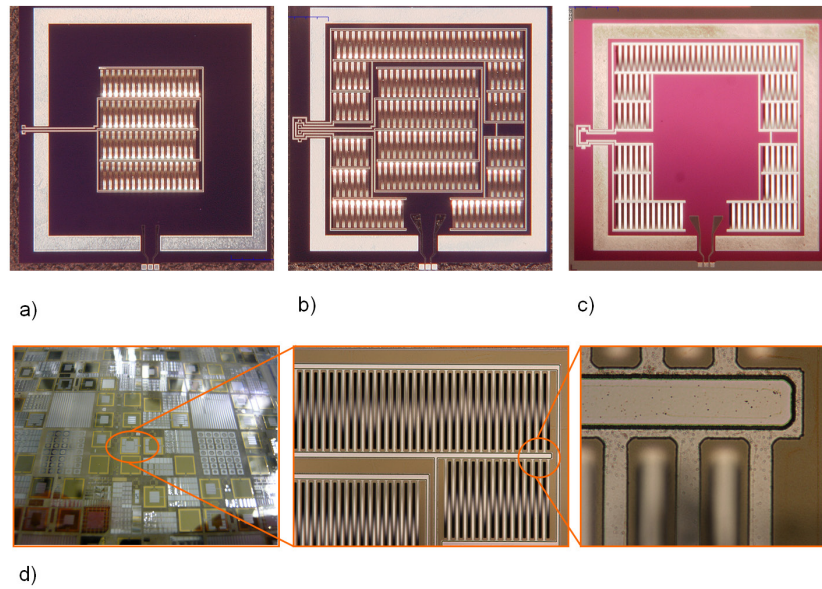


Fig. 3.38.: Realization of the FAKIR run. Volumes of perturbation: a) CPP, b) both CPP and SAP and c) SAP. d) FAKIR wafer and zoom on the MEMS.

Additionally to the variety of feeds, three different internal volumes of perturbation were implemented using EFFA MEMS and are presented in Figure 3.38. These were largely discussed in Section 3.2.2. As a reminder, we recognize in Figure 3.38-a) the CPP, -b) both CPP and SAP and -c) SAP. The layouts with the first feeding are presented in Figure 3.34 at locations b)-1 and f)-1 with CPP; with the second feeding at locations c)-1 and f)-2 with CPP; with the third feeding at locations d)-1 and f)-4 with CPP; with the fourth feeding at locations d)-4 and f)-3 with CPP. The fifth and sixth feeds are more largely used and all kinds of internal volume of perturbation are addressed: the fifth feeding is presented in a)-1 with mix CPP-SAP, e)-1 with CPP, e)-2 with SAP, e)-3 and e)-4 with CPP; the sixth feeding is presented in a)-2, e)-1 with CPP, b)-2 with CPP, c)-2 with SAP and d)-3 with mix CPP-SAP.

### 3.3.3.3 Integration in MCM-D: MONOCEROS run

Four cells were designed for the MONOCEROS run, but only two cells are of interest in this section. Those cells correspond to air cavity resonators of which layouts are presented in Figure 3.39. The purpose of those implementations is to assess the integration of microstrip line fed cavity resonators, similar to the ones discussed at 4 GHz, in Section 2.4.2 and single cavity filters at 60 and 77 GHz a) in a vias-less version and b) in a version where the CPW to microstrip lines require grounding TSVs. The other two cells were partially presented in Section 2.5 for the characterization of the substrate integrated cylindrical cavity resonators and the oscillators based on those substrate integrated cavity resonators will be presented in Chapter 5.

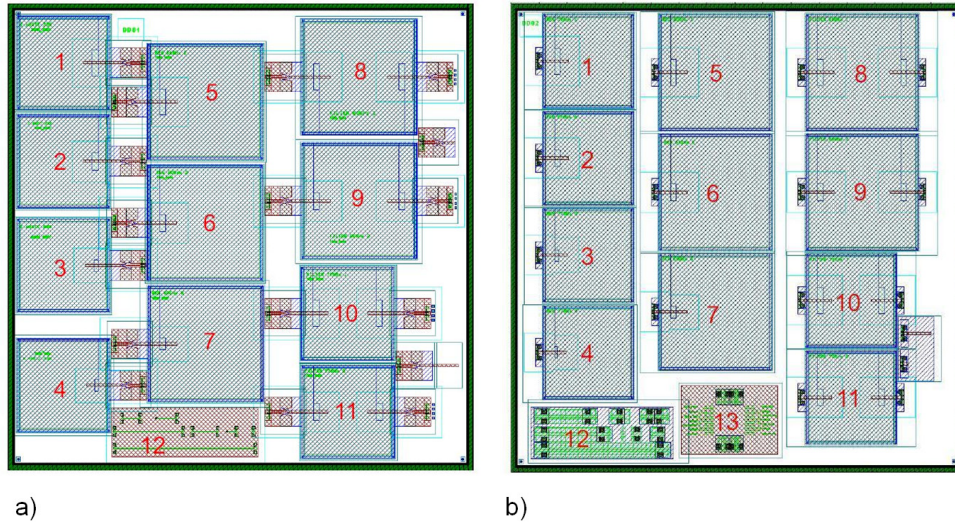


Fig. 3.39.: Layouts of the MONOCEROS run a) in a vias-less and b) with TSV microstrip feeds, incorporating twenty-two different cavity resonators.

The coupling to the cavity is realized through an aperture in the back metallization which dimensions fix the coupling coefficient, fed by a microstrip line. At 60 GHz

the dimensions of the cavity are  $3855 \times 3855 \mu m$  and the aperture dimensions are respectively: a,b)-1, overcoupled feeding,  $900 \times 200 \mu m$ ; a,b)-2, overcoupled feeding,  $600 \times 200 \mu m$  and a,b)-3, critically coupled feeding,  $400 \times 150 \mu m$ . At 77 GHz the dimensions of the cavity are  $3100 \times 3100 \mu m$  and the aperture dimensions are respectively: a,b)-1, overcoupled feeding,  $900 \times 200 \mu m$ ; a,b)-2, overcoupled feeding,  $600 \times 200 \mu m$ ; a,b)-3, critically coupled feeding,  $400 \times 150 \mu m$  and a,b)-4 a second critically coupled feeding,  $400 \times 100 \mu m$ . Are also presented devices a,b)-8 with an aperture of  $900 \times 200 \mu m$  and a,b)-9 with an aperture of  $600 \times 200 \mu m$ , which are air cavity filters at 60 GHz and devices a,b)-10 with an aperture of  $900 \times 200 \mu m$  and a,b)-11 with an aperture of  $600 \times 200 \mu m$ , which are air cavity filters at 77 GHz.

### 3.3.4 MEMS design

The EFFA or Electrostatic Fringing-Field Actuators introduced by [Rottenberg 07], consist of two sets of interdigitated cantilever beams realized on an insulating substrate. In the idle state the beams curl out-of-plane due to their residual stress gradient as shown in Figure 3.40-a). A particular MEMS derived is adapted to be integrated in the cavity resonator. The main difference is that a bottom electrode is isolated from the two sets of interdigitated cantilever beams by a dielectric layer.

In the absence of bottom electrode, the EFFA actuates in two steps. First, a DC-voltage applied between the two electrodes produces fringing electric fields that result in opposite bending moments moving downwards both sets of fingers. The fingers are coming closer to the insulating substrate, characterized by a permittivity  $\epsilon_r > 1$  and this proximity produces an unbalance of the E-field distribution between top and bottom faces of the fingers. The electric field lines close mainly through the substrate. A biased virtual electrode is located below each finger that applies thus on these a net downward force. As a result, the fingers tend to contact the substrate, while remaining aligned with each other. For the proper actuation of the fingers, the design relies on the lateral rigidity of the thin film cantilever beams preventing the side pull-in [Rottenberg 07].

The presence of the bottom electrode, which is inherent to the cavity resonator where it maintains the fields integrity, is changing the actuation mechanism. Actually the actuation between the grounded bottom metallization and the bias on the beams is dominating the actuation. This is visible on the mechanical simulations with the software COVENTORWARE [Coventor], presented in Figure 3.40. The simulations were performed considering the measured initial deflection  $\delta_{EFFA} = 30 \mu m$  of the cantilevers with a length,  $L_{EFFA} = 500 \mu m$ , based on which, the strain gradient is calculated:  $\Gamma_{EFFA} = 2\delta_{EFFA}/L_{EFFA}^2 = 240 m^{-1}$ . Considering the strain of the AlCu<sub>0.5%</sub> alloy is  $E_{AlCu_{0.5\%}} = 71$  (MPa), a residual positive stress gradient of  $17 \text{ MPa} / \mu m$  is used with the simulator. The result of the simulations is that the pull-in voltage is smaller in the case of an even actuation i.e. when the voltages applied on both sides of the array of beams is the same:  $V_1 = V_2$ , versus a differential actuation when  $V_1 = -V_2$  also applied at each sides of the array of beams, behavior which is confirmed by the measurements. It is also demonstrated that the beams can be hold down by using bipolar actuation with switching actuation voltages on both sides, no stiction is resulting when beams are hold down for a long period.



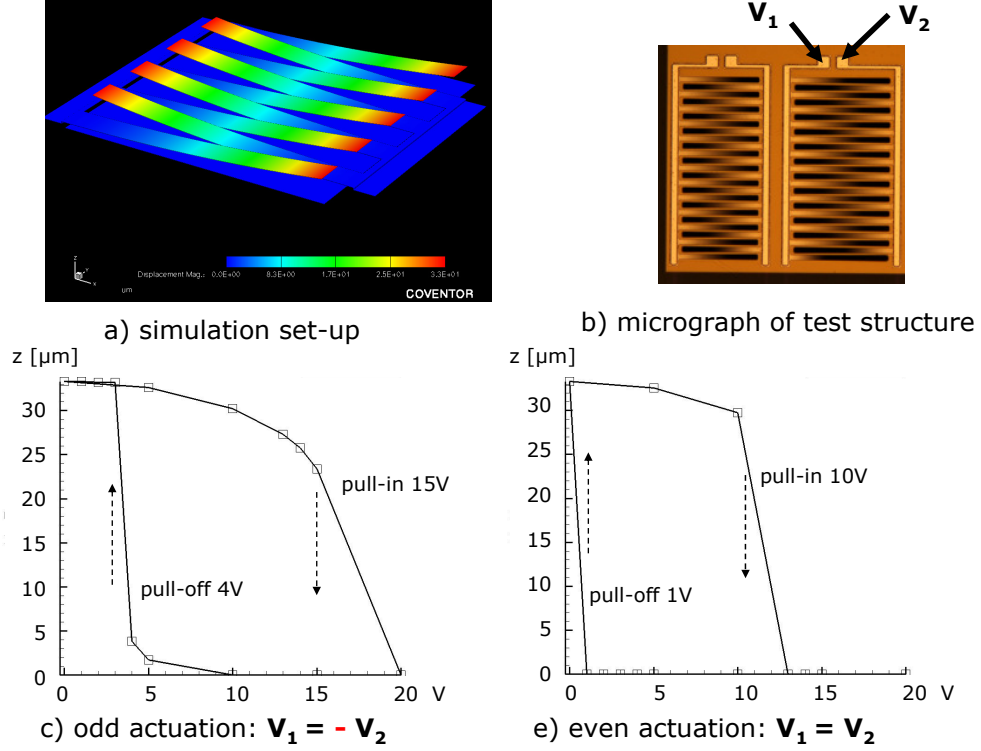


Fig. 3.40.: a) COVENTORWARE's simulation set-up of the MEMS actuation, b) micrograph of test structure with actuation probes applying d) a differential DC voltage e) the same voltage on both sides of the cantilevers array.

#### 3.3.4.1 MEMS characterization

During the measurements the capacitance is evaluated given a DC potential is applied between the two arms of the cantilevers presented in Figure 3.40-b). A double voltage sweep from 0 to 40 V and back from 40 to 0 V, with a step size of 0.5 V is applied. The impedance is measured at each voltage step and the capacitance is extracted, see Figure 3.41.

From the C-V plots the pull-in and pull-out voltage,  $V_{PI}$  and  $V_{PO}$  respectively, become visible and are extracted. For the rather stiff cantilevers from 3.41-a), it is found that the pull-in voltage  $V_{PI} = 31$  V and the pull-out voltage  $V_{PO} = 0$  V. Beyond the pull-in voltage ( $V > V_{PI}$ ) the capacitance continues to rise with increasing voltage. This can be explained by a continuous narrowing of the air gap at the anchoring of the cantilevers. If the graph is inverted an idea of the relative deflection of the EFFA cantilevers can be obtained. At the pull-in voltage the cantilevers are close to the substrate and at the pull-out voltage they are back in their initial curled out-of-plane position. A parametric study of the EFFAs was performed to identify the structures/designs that were not prone to stiction after the first actuation. Measurements were made of

169 out of the 210 different EFFA cantilever array designs as follows: lengths of 400, 500, 600, 700, 900, 1100 and 1200  $\mu\text{m}$ , widths of 20, 25, 30, 40 and 50  $\mu\text{m}$  and slots of 2, 3, 4, 5, 7 and 9  $\mu\text{m}$ . Identification of the structures returning to their curled-up position was based on comparison of the value of the capacitance of the MEMS before ( $C_{start}$ ) and after the actuation cycle ( $C_{end}$ ). If these were equal, as is the case for the device in Figure 3.41-a), the cantilevers did not stick and returned to their original position. If on the other hand  $C_{start} \neq C_{end}$  (in fact  $C_{end} > C_{start}$ ), as is the case for the compliant cantilevers of Figure 3.41-b), the cantilevers remained stuck upon releasing the voltage.

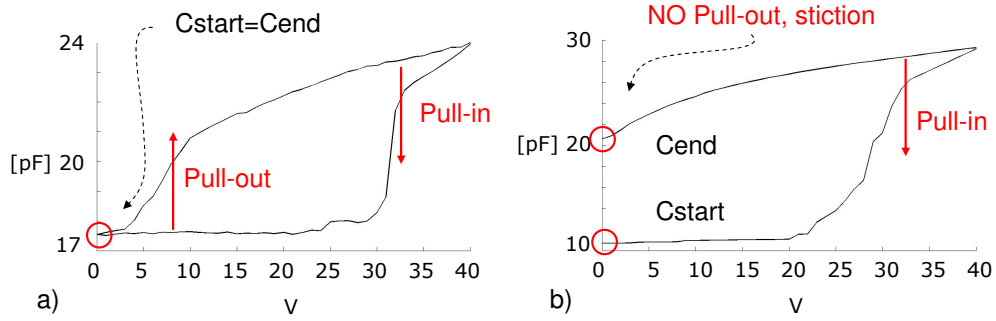


Fig. 3.41.: Measured C-V curves of EFFA MEMS. a) For rather stiff cantilevers: 700  $\mu\text{m}$  long, 50  $\mu\text{m}$  wide and 5  $\mu\text{m}$  slot. b) For rather compliant cantilevers: 1100  $\mu\text{m}$  long, 20  $\mu\text{m}$  wide and 5  $\mu\text{m}$  slot.

The results are summarized in Figure 3.42 showing the measured  $C_{end}$  vs.  $C_{start}$ . The reliable cantilever arrays are located close to the line representing equality of  $C_{start}$  and  $C_{end}$ . The measured C-V characteristics of the EFFA cantilever arrays that are reliable provide the same values of capacitances before ( $C_{start}$ ) and after ( $C_{end}$ ) actuation were observed, which is an indication that the cantilevers properly released upon removal of the actuation voltage. It is observed that the width, and the length (but to a much lower degree), seems to be the determining factor in attaining reliable behavior from the EFFA cantilevers. Cantilevers with a width of 50  $\mu\text{m}$ , compared to 20  $\mu\text{m}$  for the current design used in the FAKIR run, demonstrated reliable behavior for lengths ranging from 500 to 900  $\mu\text{m}$  and spacings from 3 to 9  $\mu\text{m}$ .

Unfortunately, the EFFA arrays implemented first inside the resonators were short-circuited which is not the case in the test array used for the parametric study. Further investigations of the process-flow has demonstrated that the shorts are between the two metallization and that they are a consequence of non-uniformly plasma loading. During the Polyimide etch, non uniform plasma loading is present around the openings in the TSM. Nevertheless, tests revealed that these cantilever designs for the FAKIR flow/design: 400-500  $\mu\text{m}$  long, 20  $\mu\text{m}$  wide and 5  $\mu\text{m}$  slot can be actuated but remain stuck after actuation. They were however used as a "one-shot device" to prove the concept, once the problem of the electrical shorts was solved.

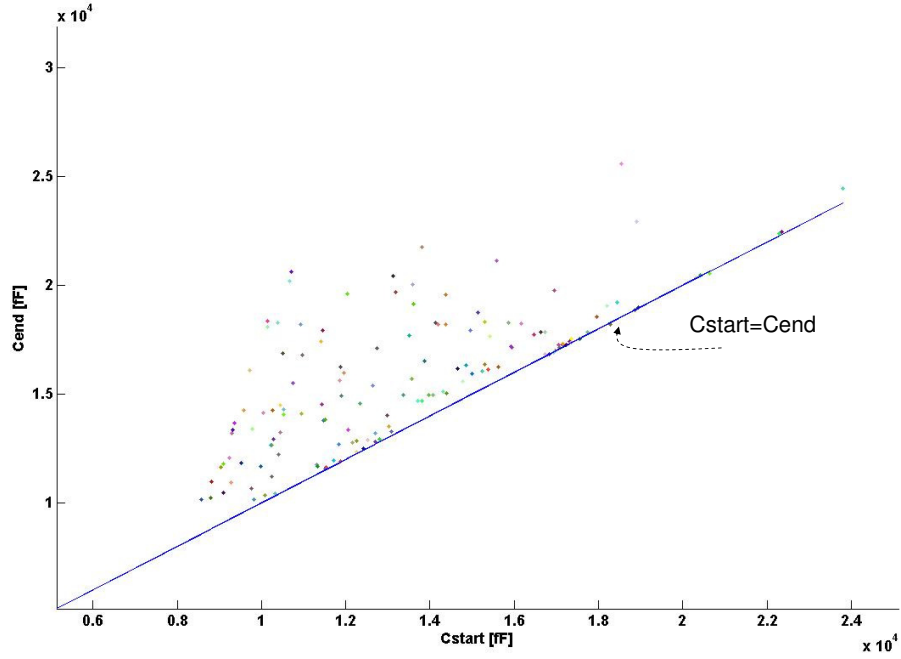


Fig. 3.42.: Classification of all the measured EFFA cantilever arrays in a  $C_{end}$  vs.  $C_{start}$  graph. Each point represents a design of an EFFA cantilever array. The straight line indicates the points where  $C_{start}=C_{end}$ .

### 3.3.5 0-level packaging of the MEMS

The standard solder (Sn) bonding procedure both carrier wafer (with the TSM, e.g. Cu(2  $\mu m$ ) / Ni(3  $\mu m$ ) / Au(0.2  $\mu m$ ) finish) and cavity chip (with the solder, e.g. CuSn) are held at 300°C during flip-chip assembly. The aluminum (AlCu<sub>0.5%</sub>) MEMS cantilevers showed significant deformation and could not be actuated after assembly under these conditions. The temperature at which plastic deformation appears for Aluminum is 100°C. The low temperature process should prevent the AlCu<sub>0.5%</sub> cantilevers to deform plastically. Key in the low temperature process is that the cavity (with the solder) is at a higher temperature (allowing the solder to melt) while the carrier chip (with the cantilevers) is maintained at a lower temperature. Quite some effort has been put in the cavity assembly on the MEMS (EFFA-based) tuneable platform.

### 3.3.5.1 *Ag-glue assembly*

To reduce the temperature exposure of the MEMS during capping, thirteen cavities equipped with a CuSn ring were assembled with Loctite 3880 epoxy glue and compressed for 300 s with 500 gF. Both carrier and cavity chips were maintained at 100°C for curing purposes. Two assemblies showed no resonance. Six out of the eleven remaining resonators displayed a frequency shift during the DC voltage sweep. This result was very encouraging as it was the first experimental proof of the tunability concept based on MEMS located inside a cavity resonator. On average, for the 60 GHz cavities the frequency shift was about 150 MHz with a maximum shift of 180 MHz (i.e. 0.3%). The unloaded extracted Q-factor  $Q_0$  was on average low,  $Q_0 \approx 35$  (maximum  $Q_0$  of 55). As the frequency tuning range was quite small, smaller than expected, and occurred for rather low voltages, typically around 8 V, compared with 15 V pull-in before capping, this suggested that the physical properties or the shape of the MEMS cantilevers were changed during the assembly process. It was further confirmed that the deflection of the cantilevers was reduced during the assembly due to the heat. Shear tests were performed on the assembled cavity resonators to assess the strength of the adhesion. During the shear test, a stress is applied horizontally to the sample until the sample fails. The shear speed is 50  $\mu\text{m} / \text{s}$  applied at a height of 50  $\mu\text{m}$ . The load force (translated in pressure) applied is recorded at the failure instant. The pressure applied during the flip-chip assembly process was either 2.6 MPa or 5.2 MPa corresponding to a force of 1 kgf or 2 kgf applied on the cap of the cavity (contact area:  $3.810^{-6}\text{m}^2$ ). The shear test results for the epoxy glue assemblies are resulting on average of a shear pressure of 34 MPa.

### 3.3.5.2 *Low temperature soldering*

An adapted assembly procedure, presented in Figure 3.43-a), was developed, whereby the carrier wafer temperature was reduced to 100°C as an attempt to prevent plastic deformation of the cantilevers. Prior to the bonding, the temperature of the cavity chip (with the Sn solder) was increased to 300°C. After alignment, the cavity chip is brought down towards the carrier wafer until contact is made. From that moment onwards a redistribution of the temperature starts to take place and the temperature at the location of the MEMS will slowly increase. The time allowed for the temperature increase is a very important parameter. The time should be long enough to ensure a sufficiently strong bond, but conversely as short as possible to prevent the temperature of the cantilever exceeding the critical temperature (which is close to 100°C for AlCu<sub>0.5%</sub> beams) above which plastic deformation occurs. Typical time intervals of 240 s were used. Unfortunately, even this low temperature soldering affected the MEMS and therefore further alternatives were investigated.

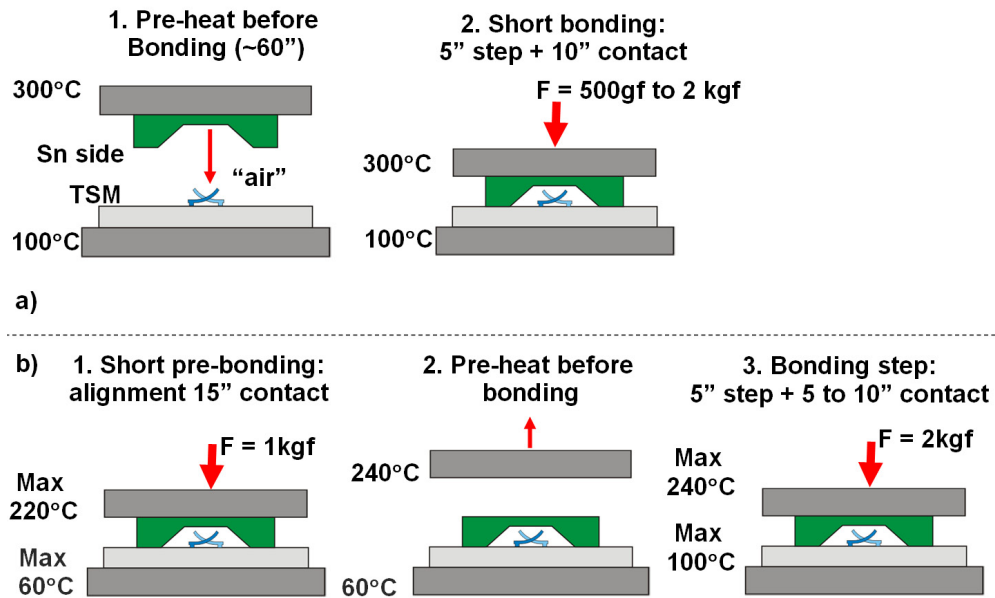


Fig. 3.43.: a) Low temperature FC process, b) improved fast soldering.

The main failure mechanism after assembly is solder squeeze-out from the TSM, effectively shorting the RF feed or biasing line. An illustration of the solder squeeze-out reflow is presented in Figure 3.44 where most of the solder is pushed out of the location of the TSM ( $\text{Cu}(2\ \mu\text{m}) / \text{Ni}(3\ \mu\text{m}) / \text{Au}(0.2\text{-}0.3\ \mu\text{m})$ ). An even further reduction of bonding time and temperature of the cavity die is expected to result in less plastic deformation on the outer ring of MEMS actuators. In total seven cavities were assembled, amongst which four showed a resonance and two have demonstrated tunability. Reduction of the bonding force and temperature is expected to reduce the risk of solder squeeze-out and increase the yield.

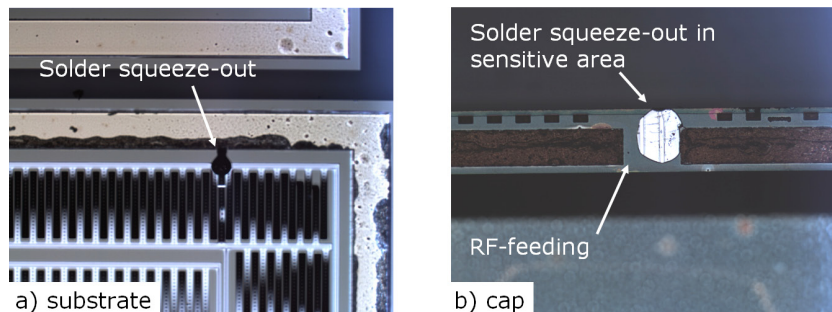


Fig. 3.44.: Micrograph of solder squeezed out due to the bonding.



Additionally, Figure 3.45 shows how the MEMS cantilevers close to the ring start to show plastic deformation after ten seconds bonding time, while the cantilevers in the center of the cavity still have their initial deflection. The resulting bond is not very strong, but sufficient to guarantee a good electrical connection between cavity and carrier substrate. Improvement of the bond can probably be achieved by reducing the contact force since with the current pressure of 1.2 MPa a significant amount of solder squeeze-out is observed.

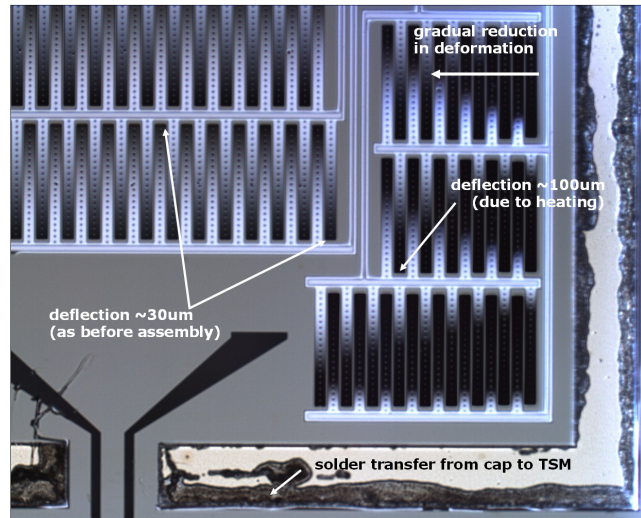


Fig. 3.45.: Micrograph of substrate, after bond and cap removal.

### 3.3.5.3 Fast and low temperature soldering procedure

<sup>1</sup> In order to reduce both the solder squeeze-out and the plastic deformation, temperature, force, soldering time as well as the operation procedure were adjusted. The main adjustments are presented in Figure 3.43-b). The alignment and positioning is performed at room temperature. The application of the cap occurs in three steps to prevent melting of the solder, before contact. After alignment a short pre-bonding first step is performed at a limited temperature of 220°C (temperature below the melting point of Sn: 231.93°C) to limit the time of oxidation of the inner metallization (Cu and Sn) and soldering squeeze out. During this step, the temperature of the MEMS is set to 60°C instead of 100°C during the final soldering step. During the second step the bonding chuck is heating without being in contact with the cavity. In the third step, the final bonding temperature is reduced from 300°C to 240°C on the cavity side to prevent/reduce solder squeeze-out and limit MEMS deformation at the outer array. Several soldering times were investigated from 10 s to 5 s. Several pressures were applied, between 1.3 MPa and 5.2 MPa, corresponding to a force of 0.5 kgf to

<sup>1</sup>This section results from a strong collaboration with P. Ekkels [Ekkels 10], within the framework of the European research project 3DμTune (IST-2005-027768).

2 kgf. The shear tests performed on the assemblies were resulting in an average shear pressures of 43 MPa, peaking at 78 MPa for the 2 kgf force and 7.5 s soldering time parameters at a temperature of 240°C. Those parameters improve both solder transfer and bond strength. Small variations of the force and time of application have a direct impact on the bonding strength while the MEMS located inside the cap preserve their integrity.

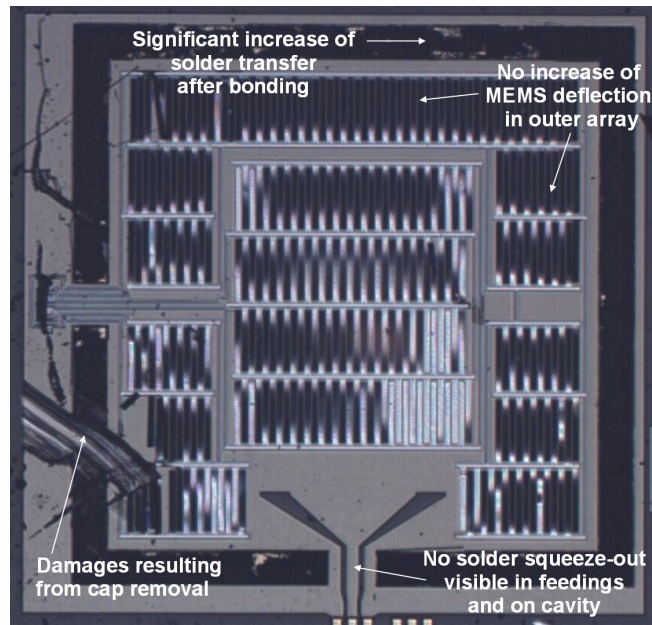


Fig. 3.46.: Micrograph resulting after the shear test of a cavity capped with the new fast and low-temperature bonding process.

### 3.3.6 Tunable cavity resonators measurements

The return loss ( $S_{11}$ ) of the resonator is measured for different DC voltages (sweep from 0 to 30 V by steps of 2 V) applied to the array of cantilevers, located in the trapezoid cavity resonator, see Figure 3.47.

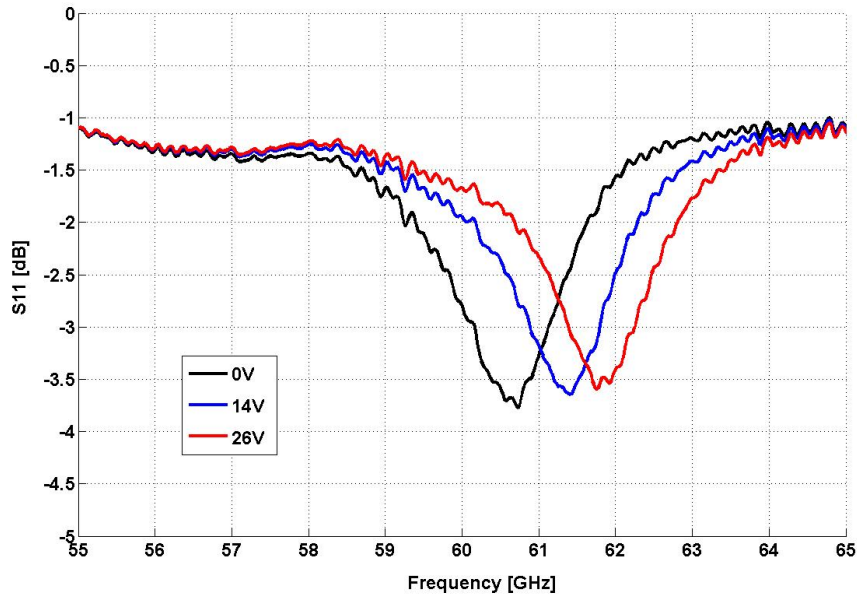


Fig. 3.47.: Measured return loss  $S_{11}$  for three different DC voltages applied to the 2D cantilever array of Figure 3.27. The minimum in  $S_{11}$  corresponds to the resonance frequency of the cavity resonator.

At each voltage step the  $S_{11}$  parameters were extracted in a RF sweep from 55 to 65 GHz. The minimum of the return loss corresponds to the resonance of the  $TE_{101}$  mode of the cavity resonator and is extracted in Figure 3.48. The shift in frequency as a function of the applied voltage  $V$  is presented in Figure 3.48. The idle resonance frequency, with no perturbation due to the MEMS, is 62 GHz. This measured frequency is close by 0.6% to the frequency of 61.6 GHz, evaluated considering the measured dimensions of the fabricated square trapezoid cavity resonator:  $a_{KOH} = d_{KOH} = 3750 \mu m$  and height  $b = 440 \mu m$ . Those measurements are inclined to some uncertainty with respect to the lateral and height dimensions. The loaded  $Q$ -factor was expected to be  $Q_l = 309$ , when loaded to a  $50 \Omega$  line ended by a feeding realized as a shorted CPW line. The expected unloaded  $Q$ -factor was simulated to be  $Q_0 = 916$  and are both extracted with QZERO from Ansoft's HFSS simulations. The extracted unloaded  $Q$ -factors from the measurements (using also QZERO), are  $Q_0 = 55$  and the loaded  $Q$ -factor is  $Q_l = 48$ . The rather low measured value of the  $Q$ -factor is attributed to a flaw in the metallization of the cavity resonator. We also measured other cavity

resonators realized in the same technology and these show much higher Q-factors (of a few hundred). Nevertheless, what is more relevant at this point is that the shift in frequency observed during the actuation of the cantilever array is in very good agreement with the theory and corresponds to the expectations. A maximal shift in frequency of 1.2 GHz (for the 60 GHz resonators) is observed for the full excursion of the cantilevers shaping the Faraday cage.

The frequency shift measured as a function of the Faraday cage height  $H_{cage}$  is presented in Figure 3.48. For a cage height,  $H_{cage} = 33 \mu m$  and lateral dimensions  $W = L = 1900 \mu m$ , a relative frequency shift of 2% is observed. The measured frequency shifts are in very good agreement with the theoretical analysis. In principle, higher tuning ranges can be achieved, provided the deflection of the cantilevers is larger.

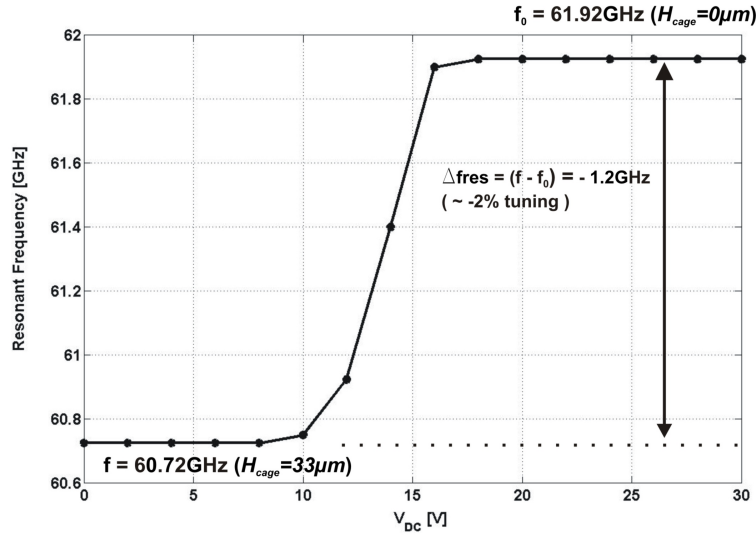


Fig. 3.48.: Resonance frequency extracted from Figure 3.47, versus the applied DC voltage (sweep 0-30 V). The frequency  $f_0$  is defined as the resonance frequency when the cage is flat,  $H_{cage} = 0 \mu m$  and  $f$ , when the cage has a volume inside the cavity resonator of height  $H_{cage} = 33 \mu m$ .

Figure 3.49 presents the measurements of the dual behavior induced by a SAP. Actuating the SAP, the resonance frequency ( $f = 61.294$  GHz) decreases at pull-in ( $f_0 = 61.286$  GHz) and stabilizes as the MEMS remain in contact with the bottom dielectric. The resonance frequency only shifts by 9 MHz (pull-in at 15 V) which is due to the actuation of only several cantilevers among the SAP array. Actually, the frequency tuning is rather low nevertheless the Faraday cage implemented by MEMS is also functional in this case. The low frequency shift measured could result from the plastic deformation of the cantilevers located closer to the soldering area. The thermal isolation from the surrounding air enclosed by the cavity is therefore reduced in com-

parison with the central area. The proof of concept of tuning by SAP is realized even though the tuning range is very small. This is resulting from the partial actuation of several cantilevers among the array of cantilevers.

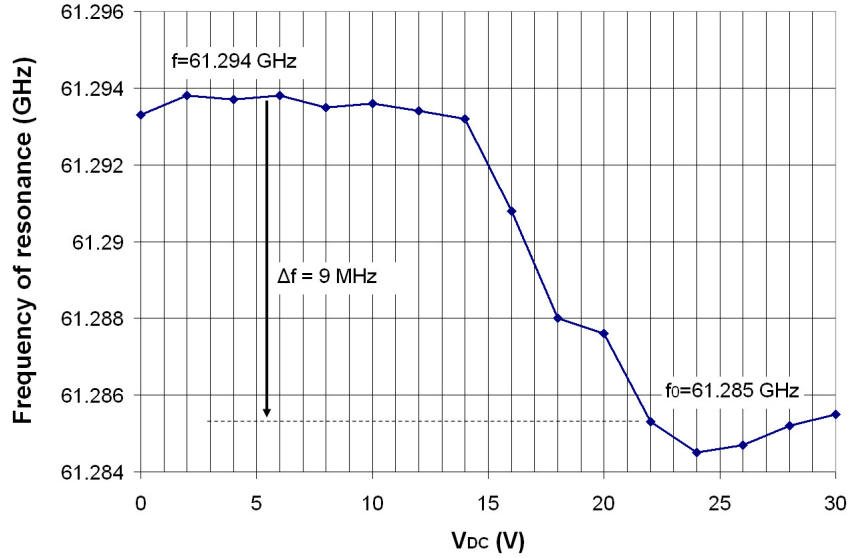


Fig. 3.49.: Resonance frequency extracted from  $S_{11}$  one port reflection measurements, versus the applied DC voltage (sweep 0-30 V). The dual behavior of the cavity resonator with SAP is demonstrated. The low the tuning range results from the partial actuation of the array of cantilevers.

### 3.3.7 Q-factor assessment

To assess the rather low Q-factor (of the order of  $Q_0 \approx 35$ ) measured during the first experiment, it was decided to assemble twelve extra cavities. Half of them by using flux during soldering and half without. The flux will dissolve any tin oxide and therefore a better metallization was expected. The six cavities with flux were subdivided into two batches. Three were assembled to form a resonator and three others were assembled above a test structure in order to measure the resistance from which an equivalent resistivity of the soldering can be extracted. Standard soldering parameters were used: bond force of 200 gF and bonding temperature of 280°C during 180 s. It was found that the mountings with flux presented a smaller bandwidth, which points towards lower losses, as compared to the mountings without flux. This observation was in agreement with the resistance measurements where the resistance across the soldering was higher on samples without flux at 1.9  $\Omega$  compared with 0.2  $\Omega$  for those with flux. All in all, on average the cavities assembled by using flux displayed a Q-factor of  $Q_0 = 50$  while for the cavities without flux a somewhat lower Q-factor ( $Q_0 = 35$ ) was obtained. These Q-values are similar to the ones from earlier experiments (e.g.

the Ag-glue mounting) and thus the reason for the low-Q could not be attributed to the presence of tin oxide.

To understand the impact of the resistive layers, a series of electromagnetic simulations realized with HFSS, were performed. It was found that a degradation of the unloaded Q-factor of one order of magnitude i.e. from  $Q_0 = 916$  to  $Q_0 = 77$  occurs when a perimetric disconnection of about  $2\ \mu\text{m}$  is considered at the edge the slanted KOH-etched cavity, as can be seen in Figure 3.50.

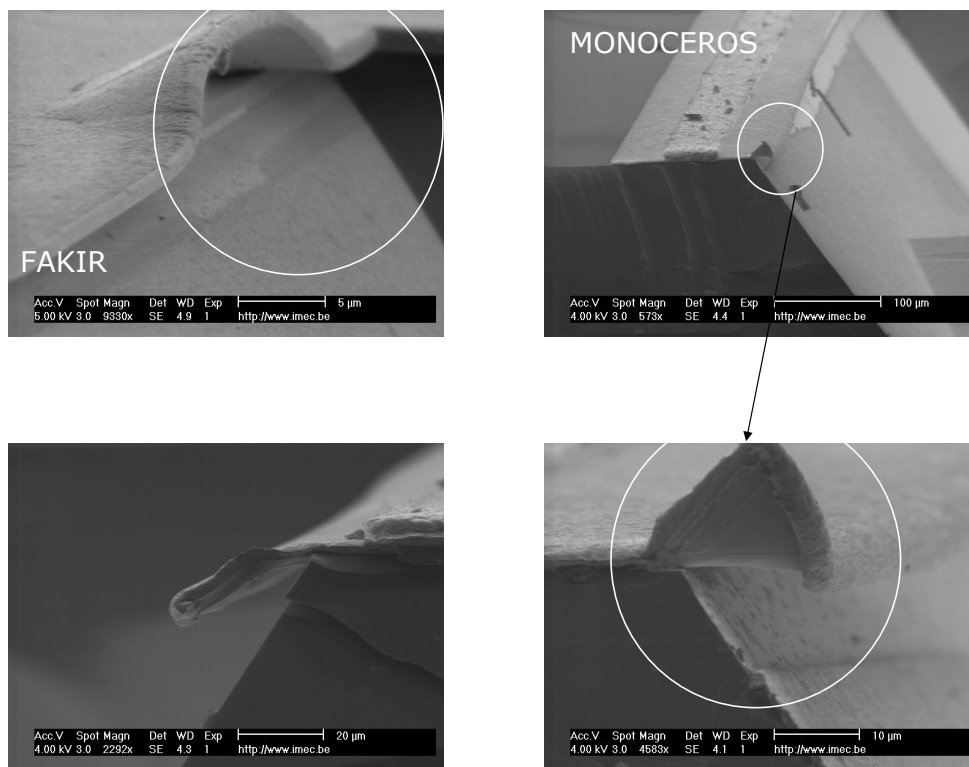


Fig. 3.50.: SEM of different cavities at the location of the flow in the metallization. The disconnection in the metallization of the cavity resonator is visible, responsible for the low Q-factors measured. SEM photos by Phillip Ekkels from IMEC, now with Epcos AG.

Further investigations revealed that the Q-factor degradation is due to a flow in the metallization of the inside of the cavity resonator. The sputter metallization results in a gutter which shadows further sputtering, disconnecting the metal layer. The gap in the metallization visible in Figure 3.50 is responsible for the high impedance band inside the cavity which consequence was simulated to reduce by a factor ten the Q-factor.



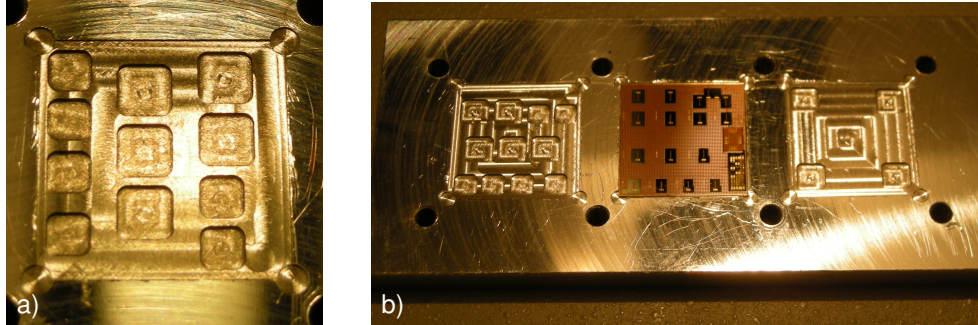


Fig. 3.51.: a) Aluminum micromachined cavities fitting the MCM-D die disposition; b) MCM-D die assembled on top of the cavities micromachined in aluminum.

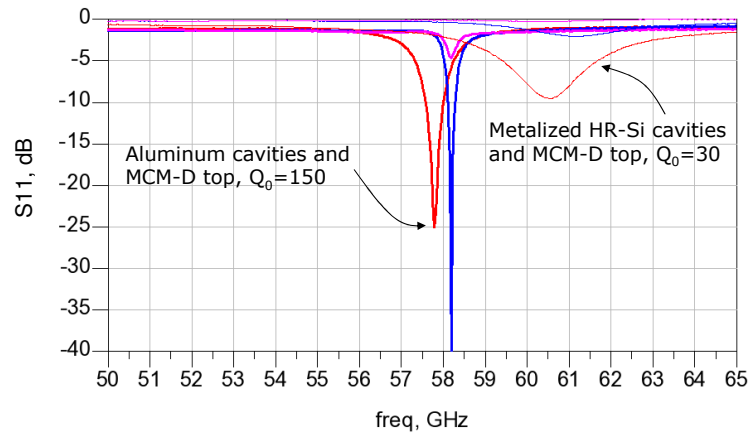


Fig. 3.52.: Comparison of the  $S_{11}$  parameter of several resonators assembled with the metallized HR-Si cavities, presenting a failure and with the cavities micromachined in aluminum.

As a consequence, new test cavities were realized rapidly using high precision machining in aluminum. Functional MCM-D dies were fixed on top of the aluminum extruded cavities. The comparison of the  $S_{11}$  parameter of several resonators assembled with the metallized HR-Si cavities, presenting a failure and with the cavities micromachined in aluminum is presented in Figure 3.52. Visible is that the Q-factor is neatly and evidently improved by a factor 5 and that the resonance frequency is shifted towards lower frequencies i.e. 58 GHz. The improvement of the Q-factor is limited essentially to the imperfect contact between the MCM-D metallization and the aluminium bloc of cavities. The frequency shift result from the imperfection of the aluminum micromachining as larger cavities were shaped.

Finally, a last experiment involved two parts not designed to fit together but well assembled by the course of events. It appeared to be the best possible option at this latter stage. The assembly is composed by a cavity designed in a preliminary run realized back to 2007, while the bottom landing surface comes from the latest Monoceros run, realized in 2010. The assembly is presented in Figure 3.53-a). This cavity was used as SEM inspection revealed that the metallization was in perfect shape, as can be see Figure 3.53-b). The final measurements and HFSS simulations,  $S_{11}$  parameters are presented in Figure 3.53-c) [Ocket 10]. The structure was designed to have a resonance frequency of 74.3 GHz and a coupling slightly overcoupled. The measured resonance frequency and unloaded  $Q$ -factor are:  $f_0 = 74.16$  GHz and  $Q_0 = 584$  while simulated:  $f_0 = 74.14$  GHz and  $Q_0 = 938$ . Actually, there is a decrease of the unloaded  $Q$ -factor due to the contact resistance, inherent to the bonding.

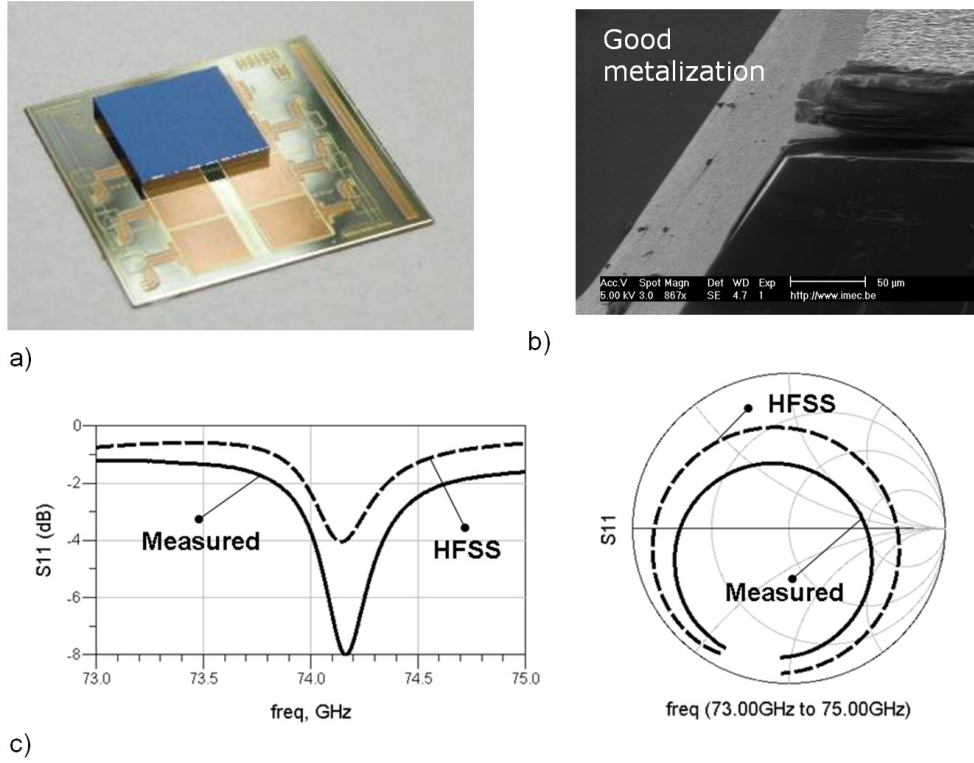


Fig. 3.53.: a) Main design made by Ocket and assessment test assembly proposed in this thesis [Ocket 10]. b) SEM of the inner cavity's metallization, no disconnection is observed. c) Measurements and HFSS simulations,  $S_{11}$  parameters of a very high performance ( $Q_0 = 584$ ) cavity resonator, at 74.16 GHz.



### 3.4 CONCLUSIONS

The 4 GHz prototyping successfully illustrated the development of the tuning theory of cavity resonators by internal non-homogeneous volumes of perturbation, i.e. CPP and SAP. The theoretical developments concerning the evaluation of the resonance frequency and Q-factor are relatively in good agreement with the measurements. However, higher than analytically calculated tuned frequencies are measured and simulated with HFSS. This is a result of the limitations of the small perturbation theory used for the analytical evaluation, as by hypothesis, such calculations are based on unperturbed fields while they are in fact well perturbed by the insertion of the perturbation volume. The impact is only rendered by finite element simulation.

It was demonstrated that both the variation of the height and the variation of the lateral dimensions of the internal volumes of perturbation have a direct impact on the resonance frequency. Using only the height variation an optimal shape of the perturbation can be identified to provide the highest impact on the frequency tuning of a cavity resonator. Using a perturbation volume composed of an array of commutable volumes of binary height (i.e. 0 or  $H_{box}$ ), the frequency could be made tunable by the commutation of different parts i.e. parts of CPP or SAP or both of them, resulting in a frequency shift of magnitude related to the location of the actuator and not only related to the amplitude of the actuation, i.e. on/off. As a consequence different zone-tuning concepts were identified and successfully implemented. The implementation concerns 72 different cavity resonators prototypes. These were simulated with HFSS (eigen mode type simulation) and measured. The full set represents a thinning range from 1 to 6 GHz from an idle resonance frequency of 4 GHz, resulting in a total tuning range of 125%. A maximal frequency shift of 3 GHz and 2 GHz is measured for respectively the resonator with CPP and SAP. Additionally, the CPP case could be used for miniaturization purposes. Cavity resonators with a resonance frequency as low as 1 GHz were measured. They are derived from an idle cavity resonator at 4 GHz. Therefore, using fixed CPP could serve miniaturization purposes.

A silicon integrated cavity resonator was made tunable, at 60 GHz by an array of cantilevers, curled upwards in the idle state and lowered down by an electrostatic actuation. The array effectively shields an air volume enclosed and therefore behaves like a Faraday cage. The shielding of the array of cantilevers is equivalent to a volume with lateral dimensions equal to the array lateral dimensions and with a height equal to the tip deflection of the cantilevers. By controlling the cantilever tip deflection with a DC tuning voltage, the cage height is effectively varied resulting in a shift of the resonance frequency of the cavity resonator. This concept was successfully used to tune a 60 GHz silicon micromachined cavity resonator by 2%. We measured a maximal frequency tuning of 1.2 GHz at 60 GHz with a tuning voltage of 16 V. This has to be compared to the first tuning results of 180 MHz attributed to the plastic deformation of the aluminum cantilevers during the curing of the epoxy glue. The proof of concept of a tuning by the external actuation zones was also realized but the tuning range was much more lower, resulting from excessive plastic deformation of the capped aluminum cantilevers. To alleviate the constraints on the bonding temperature, alternative materials as platinum and tungsten could be used to shape the MEMS.

Therefore, one of the main results of this PhD are the proof of concept of the tunability by both CPP and SAP arrays implemented with MEMS. Nevertheless, the majority of measurements were unsatisfactory regarding the expected Q-factors. The feeding structures could only partially be investigated but different coupling coefficients were demonstrated. The degradation of the Q-factor is due to a discontinuity of the metallization of the inner cavity resonator appearing during the sputtering of the metallization. This is due to the shadowing of a sputtered metallization, at the slanted edges of the cavity resonator. As a result of this investigation, a very high Q-factor resonator was finally measured at 74 GHz.

Further research is required to achieve the realization of an integrated air resonant cavities presenting both high Q-factors and high (reliable) tunability. Despite this, the tuning concept was experimentally proven and separately very high Q-factors resonators were demonstrated. The issues encountered were understood and turned into a valuable know-how towards the realization of such resonators, meeting the rigorous requirements imposed by the fabrication of such components.

## CHAPTER 4

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# PASSIVE APPLICATION: LARGE BANDWIDTH SEVEN-POLE FILTER

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### 4.1 INTRODUCTION

A microwave filter is a two-port network used to control the frequency response at a certain point in a microwave system by providing transmission at frequencies within the passband of the filter and attenuation in the stopband of the filter [Pozar 05]. Theory and practice concerning filter design stems from before World War II by Mason, Sykes, Darlington, Fano, Lawson and Richards [Pozar 05]. Reference work gathered since 1950s, mentioning are *"Microwave Filters, Impedance-Matching Networks, and Coupling Structures"* by Matthaei, Young and Jones [Matthaei 80] and the *"Foundations for Microwave Engineering"* by Collin [Collin 66], which are still of great value today. Since then, the subject has been extensively covered, for example *"Microwave Engineering"* by Pozar [Pozar 05], as a first source of information and design. The initial design, resulting from the analytical developments gathered in such references, needs to be placed in perspective towards realization and adapted to the constraints set by the technology. Actually, finite element simulations are required to cope with the complexity of 3D structures and we have extensively used HFSS for design purposes, for optimization and finally to assess the technology [HFSS ].

The spectrum in the 57-64 GHz frequency range is large and unlicensed, additionally high transmission power is allowed. Therefore, the large bandwidth per channel available enables less bps/Hz and therefore reduces the complexity of the modulation towards lower cost systems and multi-gigabit wireless transmissions [Caetano 05]. These wireless communications are promising applications as video downloads, high debit connections between laptops, beamers and printers. In this context, different

kinds of filters were demonstrated for the 60 GHz band [Lee 06, Seki 06] as strip line and microstrip line structures based on Low Temperature Co-fired Ceramic (LTCC). The drawback of these components is their high insertion loss ( $IL$ ). Alternatively, Substrate Integrated Waveguide (SIW) filters were demonstrating improved insertion loss ( $IL < 3$  dB) size reduction, integration and high performance [Lee 05a, Maruhashi 05]. However, these filters were quite narrow band, apart from [Lee 09], and we foresee the necessity of isolating the entirety of unlicensed band from adjacent frequencies, e.g. RADAR at 77 GHz, with the outcome of increasing applications.

The shift towards higher frequencies in wireless communications i.e. 40 to 80 GHz revealed the potential of integrated waveguide based solutions as outperforming other lumped passive devices, which become limited at high frequencies [Tilmans 03]. Additionally, requirements are identified in terms of weight, size and cost reduction, packaging and high density integration. Several solutions are under development, among which we investigated integrated cavity resonators in MCM-D, presented in Chapter 2. Similar characteristics are presented by the LTCC technology. This technology is composed of a multilayered platform that enables the integration of passive components and the mounting of components. It has similar RF functionalities as the MCM-D, but presents different electric characteristics and is technologically very different.

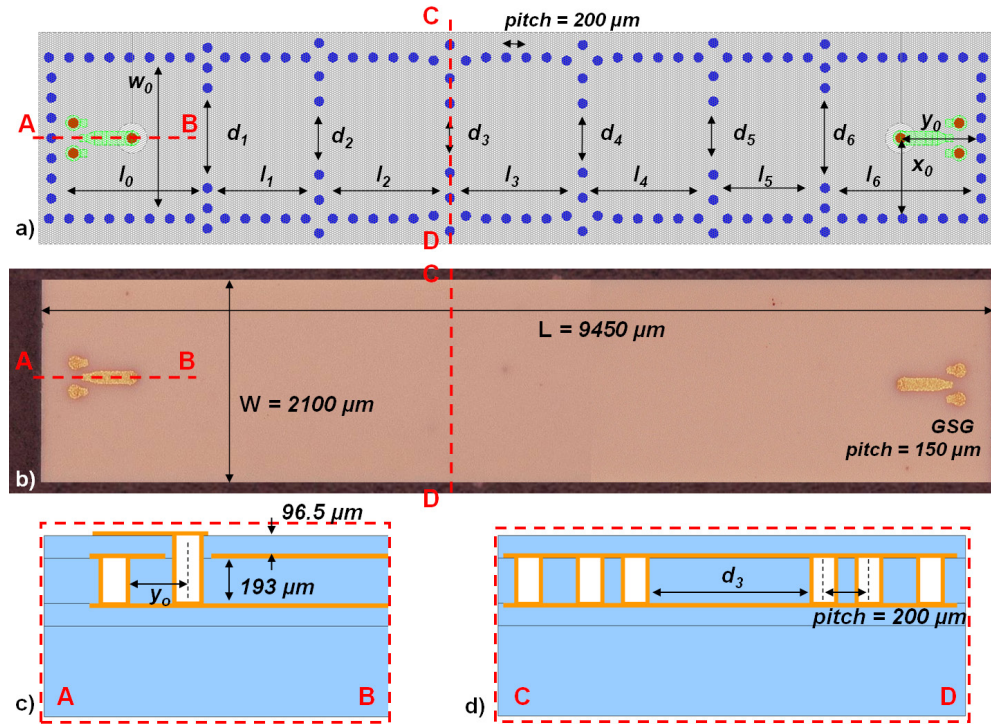


Fig. 4.1.: a) Filter parameters definition and layout, b) realization in LTCC, c) cross-section AB and d) cross-section CD.

Therefore, the focus is set towards the realization of a SIW filter in LTCC with a large bandwidth, covering the full unlicensed spectrum around 60 GHz, while offering sharp stopband attenuation. These characteristics can be achieved by increasing the number of poles of a filter [Pozar 05]. Actually, a seven-pole substrate integrated waveguide filter in LTCC is designed and realized in this Chapter. This is realized as a passive application of cavity resonators, of which theory was previously developed in Chapter 2. The filter is schematically illustrated in Figure 4.1-a) and final realization in Figure 4.1-b). After optimization, the filter was realized by the Technical research center of Finland [VTT]. Layers of A6-M material of standard thickness 5-10 mil (127-254  $\mu m$ ) are stacked and sintered to form the LTCC [Ferro]. The relative permittivity of the substrate is  $\epsilon_r = 5.7 - 6.1$  and the substrate loss is  $\tan \delta < 0.002$ . Gold based conductors 10  $\mu m$  thick are used. Vias of 100  $\mu m$  diameter are available with a minimum via to via pitch (center to center) of 200  $\mu m$ .

## 4.2 TECHNOLOGY

The filters prototypes are realized by the Technical Research Center of Finland, VTT in Low Temperature Co-Fired Ceramic (LTCC) [VTT]. The relative permittivity of the substrates are  $\epsilon_{LTCC} = 5.7 - 6.1$  and  $\tan \delta < 0.002$  vs.  $\epsilon_{HR-Si} = 11.9, \tan \delta = 0.0026$  for the HR-Si of the MCM-D. Actually, the electric characteristics belong to the A6-M material, supplied by Ferro [Ferro], layered tapes which stacking form the LTCC. The tapes are of standard thickness 5-10 mil (127-254  $\mu m$ ) and 10  $\mu m$  thick gold-based conductors are used. The variations in dielectric properties are measured and certified [Kondratyev 10].

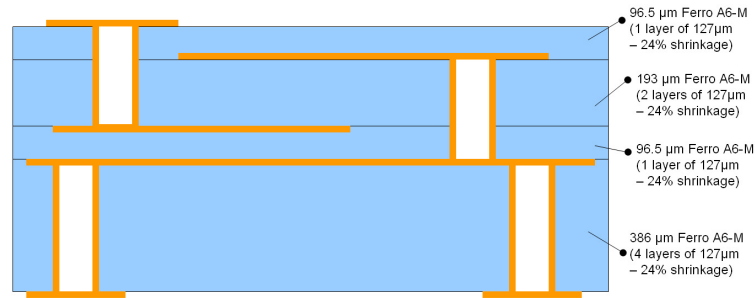


Fig. 4.2.: Cross section of the LTCC run with layer definition and characteristic thicknesses.

The typical process flow starts with the thin unfired tape sheet layers on which the conductors are shaped using conventional thick film printing. Interconnection through tapes is made using vias which are punched holes filled with conductive paste. The layers are next aligned and laminated at 21 MPa and 70<sup>0</sup>C during 5-10 minutes. The sintering step consists of a burnout at 450<sup>0</sup>C for 2 hours, with 2<sup>0</sup>C/min ramp and firing at 850<sup>0</sup>C peak for 10 min [Ferro 07]. The sintering makes the particles adhere to each other by heating the material below its melting point. The layer build-up,

as defined for the run where the cavity resonators at 60 GHz were incorporated is presented in Figure 4.2. Visible are six layers of A6-M, with corresponding vias of 100  $\mu\text{m}$  diameter size. The minimum via to via pitch (center to center) within the same tape is 200  $\mu\text{m}$ . The stacking of vias is possible through any number of layers and the top layer is a bondable metallization. The fabrication tolerances of LTCC technology allows 100/100  $\mu\text{m}$  features and spacings with 10 % accuracy [VTT 08]. Etching tolerances of metallizations and alignment tolerances of different layers are acceptable, as very good results were obtained with the integrated cavity resonators. Additionally the shrinkage is reproducible and well controlled, with the tolerance of circuit features in x and y directions, typically less than  $\pm 0.1$  % and the layer thickness tolerance is  $\pm 7$  % [VTT 08].

### 4.3 FILTER DESIGN

The insertion loss filter synthesis method starts with the specification of a physically realizable frequency characteristic (i.e. the transfer function), from which a suitable filter network is realized [Collin 66]. From the transfer function, the input impedance and the complex zeros and poles are found by various algebraic procedures, such as partial-fraction expansion [Pozar 05]. The element values of the circuit are derived from the expansion of the input impedance and serve to build the lumped element filter presenting the desired frequency response. The specification also involves the response of the filter in the passband as the Maximally Flat and the Chebyshev characteristics. The latter, allows a greater roll-off beyond the cutoff frequency at the expense of equal magnitude ripples in the passband. The Chebyshev's polynomials are tabulated and well documented in the literature [Matthaei 80]-[Collin 66]. The synthesis begins with low-pass filter prototypes normalized in terms of frequency and impedance. Then, appropriate transformations are applied to convert the low-pass design to the desired filter operating frequency and impedance level. In microwave filter design it is convenient to use a modified prototype that consists of identical series resonators coupled by impedance inverters [Harle 03]. Therefore, three important steps can be identified in the design of the waveguide filter.

Firstly, based on the specifications of the filter, the tabulated element values of the prototype filter provide values for the coupling coefficients between the resonators composing the filter as well as the value of the external coupling (the external Q-factor  $Q_e$ ).

Secondly, the synthesis continues with the low-pass filter prototype normalized in terms of frequency and impedance. Then appropriate transformations are applied to convert the low-pass filter in to the desired operating frequencies and impedance level. Thereafter the coupling coefficients are translated into physical dimensions given the waveguide structure and following the extraction of electrical parameters.

Finally, the full wave simulation is set-up and the physical dimensions are optimized.

### 4.3.1 Filter-synthesis

Firstly, we determine the order of the Chebyshev filter by computing the  $IL$  versus the normalized frequency as a function of the number of poles ( $N$ ), following [Pozar 05]:

$$IL \simeq 10 \log \left( \frac{k^2}{4} \left( \frac{2\omega}{\omega_c} \right)^{2N} \right) \quad (4.1)$$

where  $k = 10^{A_p/10} - 1$ . Considering an equal ripple attenuation,  $A_p = 0.0432$  dB corresponding to a return loss ( $RL$ ) of 20 dB in the passband i.e. 57-64 GHz (bandwidth,  $BW = 7$  GHz), centered at  $f_c = 60.5$  GHz (fractional bandwidth,  $FBW = 11.6\%$ ) and high stop band attenuation. It results that with  $N = 7$  poles, the  $IL$  is already -17 dB at 56 and 65 GHz.

The filter is composed by seven cavity resonators placed in a row, each being coupled by irises of opening length,  $d_i$  opened in their common wall. The irises can be modelled by a shunt susceptance  $B_i$  of which design formulas are outlined here below, following Collin [Collin 66]:

$$B_1 = \frac{1 - \omega/g_1}{\sqrt{\omega/g_1}} \quad (4.2)$$

$$B_2 = \frac{1}{\omega} \left( 1 - \frac{\omega^2}{g_1 g_2} \right) \sqrt{g_1 g_2} \quad (4.3)$$

$$B_k = \frac{1}{\omega} \left( 1 - \frac{\omega^2}{g_k g_{k-1}} \right) \sqrt{g_k g_{k-1}} \quad (4.4)$$

$$B_N = \frac{1 - \omega R/g_{N-1}}{\sqrt{\omega R/g_{N-1}}} \quad (4.5)$$

where

$$\omega = \frac{\pi \beta_2 - \beta_1}{2 \beta_0} \quad (4.6)$$

with  $\beta_j = \sqrt{k_j^2 - k_c^2}$ ,  $k_j = 2\pi f_j \sqrt{\varepsilon \mu}$ ,  $k_c = \pi/w_0$  and  $R = 1$  for  $N$  even and also for  $N$  odd in the case of maximally flat filters, otherwise it is given by  $R = 2k^2 + 1 - \sqrt{4k^2(1 + k^2)}$  where  $k^2$  is equal to the ripple in the passband [Collin 66]. The  $g_i$  are the element values from the low-pass prototype filter for which general solutions are provided in tables [Collin 66]. The resulting susceptances  $B_i$ , calculated using (4.2-4.5) are presented in Table 4.1.

Table 4.1.: Element values  $g_i$ , susceptance  $B_i$ , coupling coefficients  $k_{i,j}$  and physical length  $l_i$  ( $N = 7$ , Chebyshev filter 0.0432 dB equal ripple).

|              |                     |                     |                     |                     |       |
|--------------|---------------------|---------------------|---------------------|---------------------|-------|
|              | $g_0 = g_8$         | $g_1 = g_7$         | $g_2 = g_6$         | $g_3 = g_5$         | $g_4$ |
|              | 1                   | 1.008               | 1.4368              | 1.9398              | 1.622 |
|              | $B_1$               | $B_2 = B_7$         | $B_3 = B_6$         | $B_4 = B_5$         | $B_8$ |
| ( $\Omega$ ) | 1.102               | 3.128               | 4.534               | 4.842               | 1.155 |
|              | $k_{0,1} = k_{7,8}$ | $k_{1,2} = k_{6,7}$ | $k_{2,3} = k_{5,6}$ | $k_{3,4} = k_{4,5}$ |       |
| (%)          | 11.52               | 9.61                | 6.93                | 6.52                |       |
|              | $l_0 = l_6$         | $l_1$               | $l_2$               | $l_3 = l_5$         | $l_4$ |
| ( $\mu m$ )  | 1442                | 1066                | 1216                | 1257                | 1262  |

The filter's parameters are indicated on the layout presented in Figure 4.1-a). The initial design starts with the design of a squared resonator of resonance frequency,  $f_0 = 60$  GHz. Equal lateral dimensions,  $w_0 = l_0 = 1442 \mu m$  ensure the highest separation between the  $TE_{101}$  and the next higher order modes [Pozar 05]. The corresponding cutoff frequency,  $f_{off} = c/2w_0\sqrt{\epsilon_r} = 42$  GHz is far from the operation frequency [Pozar 05]. Conveniently the input and output external quality factor,  $Q_e$  can be expressed in terms of the prototype's element values,  $g_i$  as follows [Harle 03, Quan 09]:

$$Q_{e_{IN}} = \frac{g_0 g_1 f_0}{BW} \quad (4.7)$$

$$Q_{e_{OUT}} = \frac{g_N g_{N+1} f_0}{BW} \quad (4.8)$$

The internal coupling coefficients,  $k_{i,j}$  are calculated following [Harle 03]:

$$k_{i,i+1} = \frac{BW}{f_0} \sqrt{\frac{1}{g_i g_{i+1}}} \quad (4.9)$$

The coupling coefficients resulting from (4.9) are also provided in Table 4.1. Finally the electrical length of the waveguide between two irises is given in (4.10) accordingly to [Collin 66]:

$$\phi_i = \pi - \frac{1}{2} \left( \tan^{-1} \left( \frac{2}{B_{i+1}} \right) + \tan^{-1} \left( \frac{2}{B_i} \right) \right) \quad (4.10)$$



and the physical length,  $l_i$  defined in Figure 4.1-a) is given by (4.11) and is also provided in Table 4.1:

$$l_i = \frac{\phi_i \lambda_{g_0}}{\pi} \frac{1}{2} \quad (4.11)$$

where  $\lambda_{g_0} = 2\pi/\beta_0$ .

### 4.3.2 Extraction of the filter's parameters

#### 4.3.2.1 The External Q-factor

The feeding is realized by a shunting vias, as can be seen in Figure 4.1-c) and in Figure 4.3. The location of the feeding  $(x_0, y_0)$  provides the level of coupling. The closer the feeding location to the center of the resonator (i.e.  $x_0 = y_0 = w_0/2$ ), the higher the coupling coefficients  $k_{0,1} = k_{7,8}$  and lower the external quality factor,  $Q_e$ , as can be seen in Figure 4.4.

The external quality factor is extracted from HFSS simulations performed on a single resonator provided with a feed, as can be seen in Figure 4.3 [HFSS] and using the commercially available program QZERO, distributed with Kajfez's book "*Q-factor*" [Kajfez 94] for the Q-factors and coupling coefficients extraction.

A coupling coefficient,  $k_{0,1} = k_{7,8} = 11.6\%$  is obtained by locating the feeding via at  $(x_0 = w_0/2, y_0 \sim 200 \mu m)$ , following Figure 4.4. The corresponding external quality factor is  $Q_e \sim 8.7$ , considering the average measured unloaded quality factor,  $Q_0 = 124$ . However, the location of the shunt vias was fixed to the center of the feeding resonators, in order to maximize the coupling coefficient. Actually, previous measurements were characterized by lower coupling coefficients than simulated. Furthermore, the coupling coefficient is only slightly increasing after  $y_0 \sim 300 \mu m$ , see Figure 4.4. Eventually, the impact of a higher coupling coefficient is compensated during the optimization, performed in Section 4.3.3.

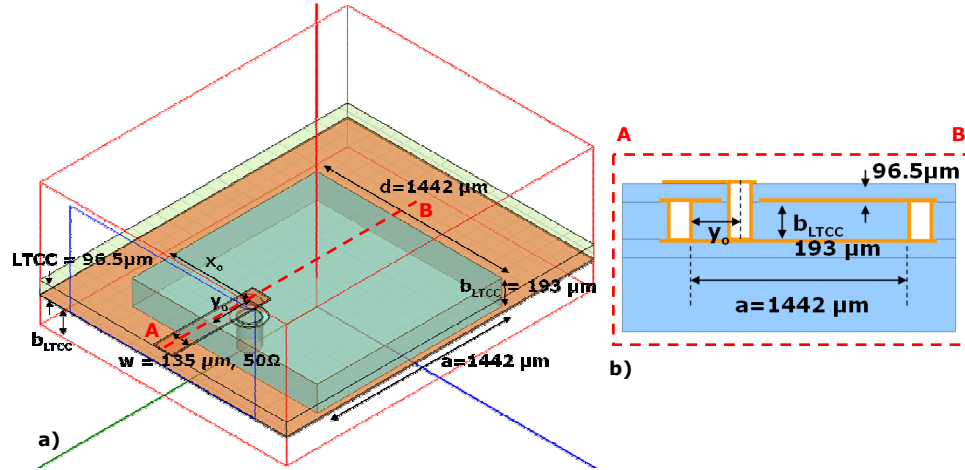


Fig. 4.3.: a) TSV feeding parameters configuration and b) transversal cross section AB, at the feed location.

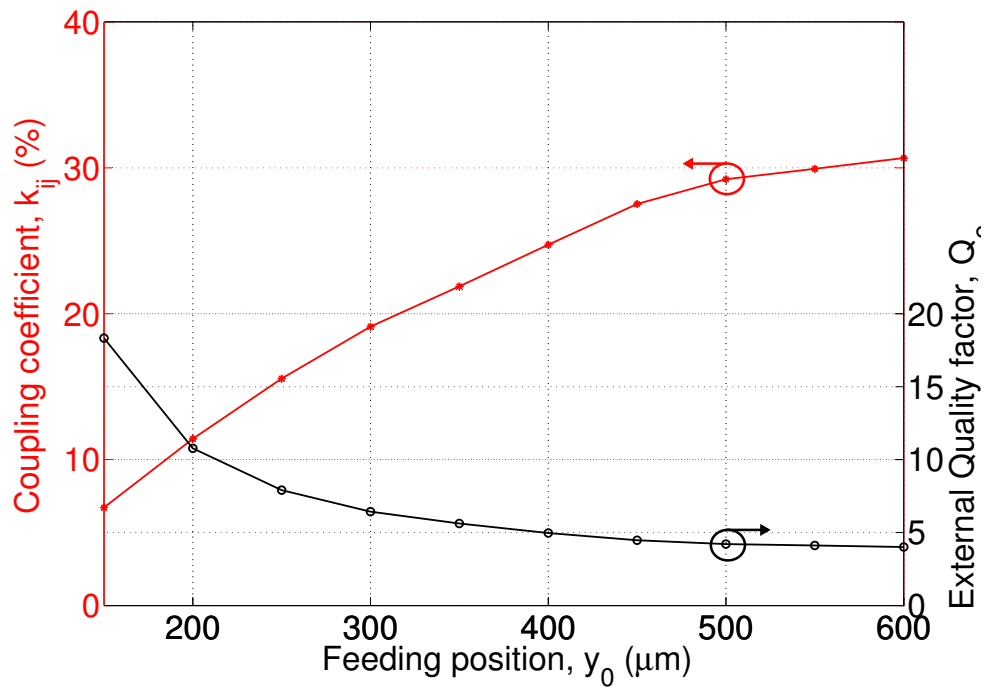


Fig. 4.4.: Coupling coefficient,  $k_{ij}$  and external quality factor,  $Q_e$  extracted from HFSS simulations with QZERO [Kajfez 94].

#### 4.3.2.2 The coupling coefficient

The coupling coefficients are extracted using the pole splitting method [Hong 96] in conjunction with full-wave simulations. In the pole splitting method, a relationship is established between the frequency separation of the poles and the coupling coefficients, as follows:

$$k_{cij} = \frac{f_{even}^2 - f_{odd}^2}{f_{even}^2 + f_{odd}^2} \quad (4.12)$$

where  $f_{even}$  and  $f_{odd}$  correspond to the frequencies of resonance of the cavity resonator when respectively a Perfect Electric, (PEC) or a Perfect Magnetic, (PMC) boundary condition cuts longitudinally through the middle of the iris between two coupled resonators. Actually, when two cavities are coupled the modes of resonance become degenerated and the resonance frequency changes according to the amplitude of the coupling. When the iris is completely opened (i.e.  $d_i = a$ ) the resulting resonance frequency is lower than in the idle case, as the lateral dimensions of the cavity resonator are larger. This corresponds to an odd field distribution of a degenerated TE<sub>101</sub> mode, that can be simulated by using a PMC boundary condition. The new resonance frequency,  $f_{odd}$  is lower than the idle one,  $f_0$  as can be seen in Figure 4.5.

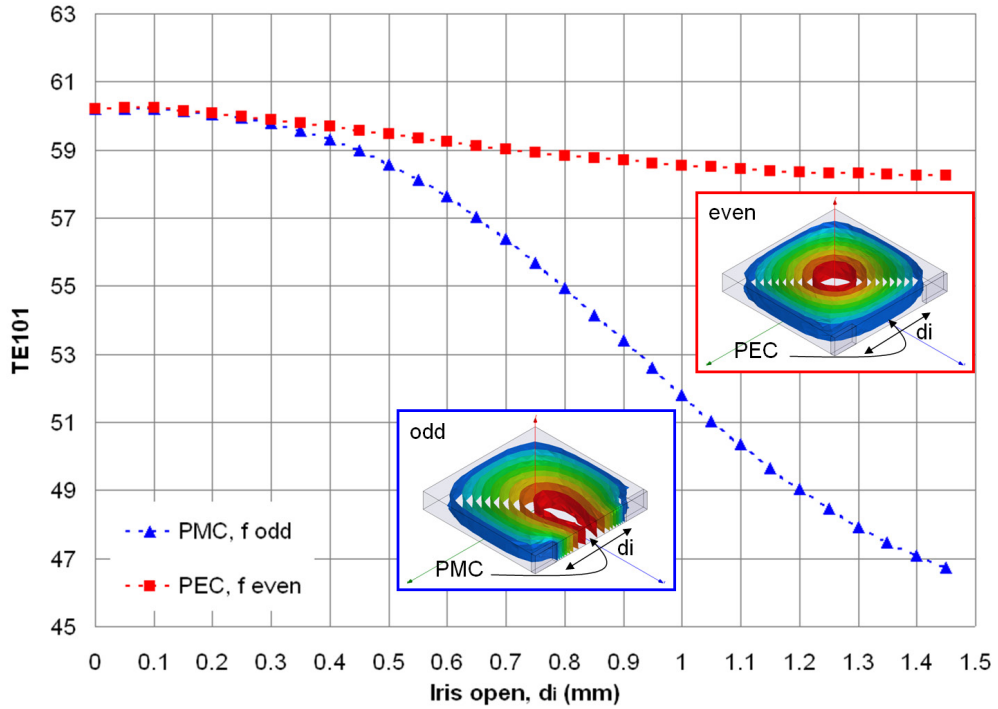


Fig. 4.5.: HFSS simulation set-up used to evaluate the coupling coefficient between two cavity resonators provided with an iris opening of length,  $d_i$ .

When the iris is completely closed (i.e.  $d_i = 0$ ), the resonance frequency,  $f_{even}$  is slightly higher than in the idle case and corresponds to a  $TE_{101}$  even field distribution and can be simulated by instead placing a PEC boundary condition. The new resonance frequency,  $f_{even}$  is higher than the idle resonance frequency,  $f_0$  as can also be seen in Figure 4.5.

As resulting from (4.12), the coupling coefficient  $k_{ij}$  is provided as a function of the iris's opening,  $d_i$ , defined in Figure 4.1-d). The values presented in Table 4.1 are extracted from Figure 4.6 and are used as first iteration of the optimization in Section 4.3.3.

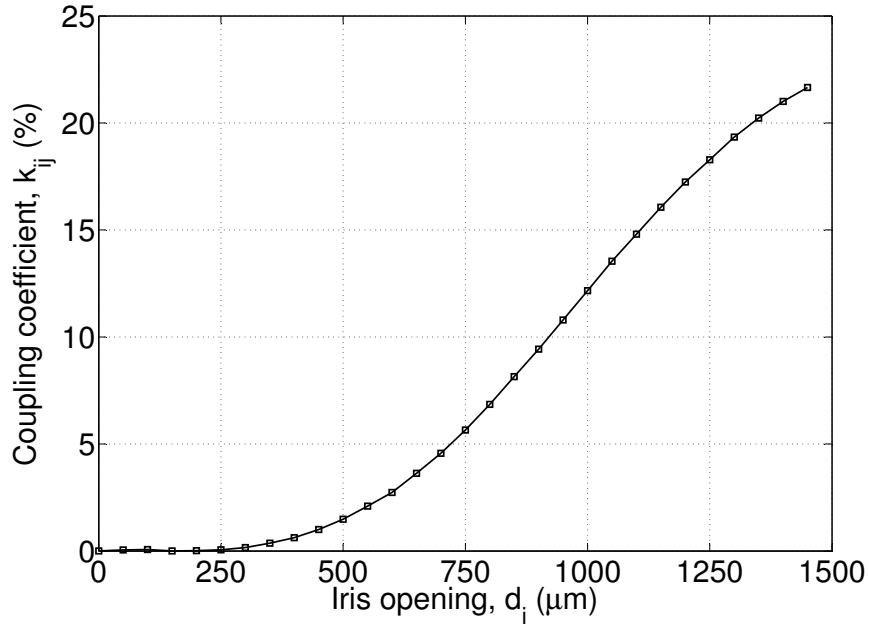


Fig. 4.6.: Coupling coefficient, evaluated using (4.12), between two cavity resonators, provided with an iris of opening length,  $d_i$ . Extraction using the pole splitting method [Hong 96].

#### 4.3.3 Simulations, optimization and measurements

The physical dimensions resulting from the filter's design, need to be adjusted when the filter is considered in its entirety. Actually, the side walls are not fully metallized, but are shaped by a wall of vias placed with a pitch of  $200 \mu m$ . Additionally, the location of the feeding vias is set to the center of the feeding resonators, providing the highest available coupling. The optimized values are slightly different and both values as well as their difference are presented in Table 4.2. The HFSS simulations

with optimized dimensions are compared to the analytical response and measurements in Figure 4.7 for the transmission and in Figure 4.8 for the return loss.

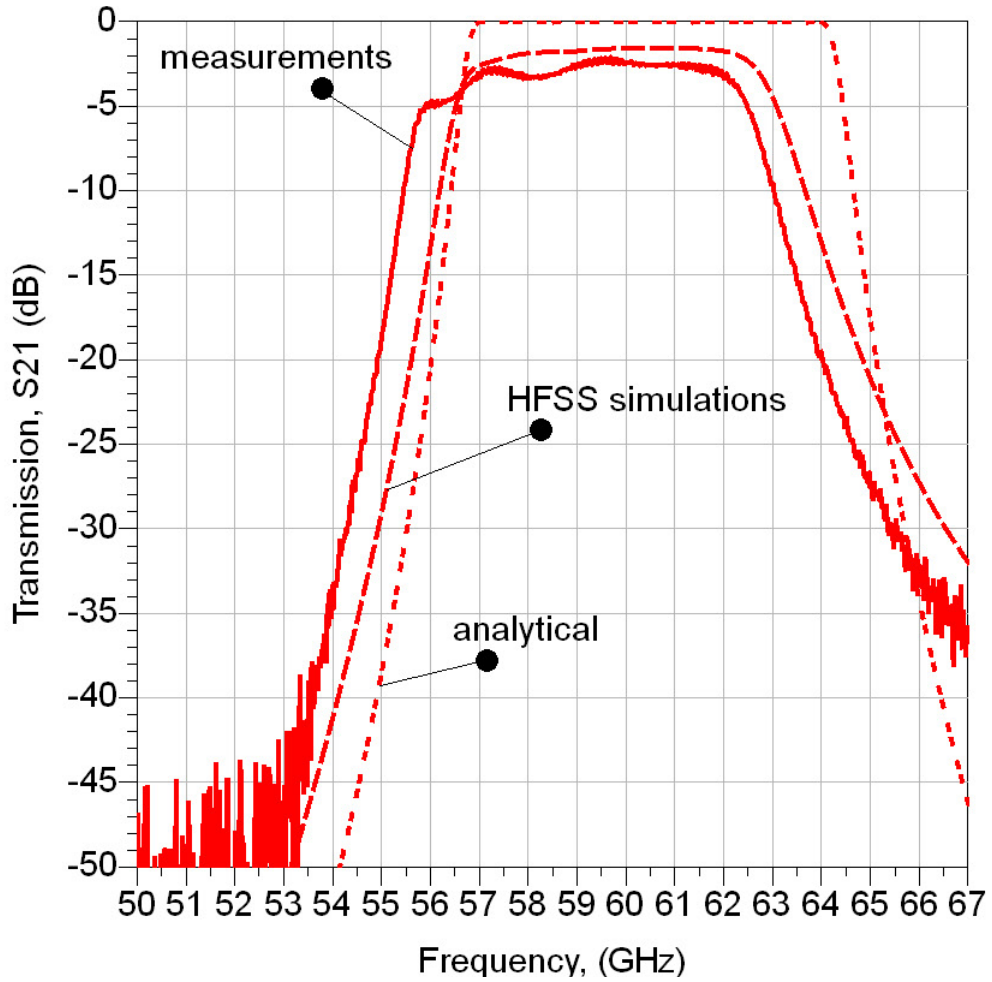


Fig. 4.7.: Comparison between measurements and simulations of the transmission parameter  $S_{12}$ .

The filter was measured using an Agilent (E8361A) Vector Network Analyzer and RF probes (1.85 mm pPROBE 67 GHz, GSG 150  $\mu m$  pitch) on a Cascade Microtec probe station. A bandwidth from 40 to 67 GHz was used for the measurements, after a SOLT calibration was performed on a separate substrate. The SOLT calibration consists in measuring a short, an open, a load, and a thru line of short length in order to mathematically remove the systematic measurement errors [Brunsman 03].

The HFSS simulation result in a bandwidth,  $BW = 6.5$  GHz (from 56.5-63 GHz),  $FBW = 10.9\%$  around a central frequency,  $f_c = 59.75$  GHz with an insertion loss,  $IL = 1.6$  dB and a return loss,  $RL < 15$  dB at  $f_c$ .

The measured bandwidth,  $BW = 7$  GHz (from 55.6-62.6 GHz),  $FBW = 11.8\%$  around a central frequency,  $f_c = 59.1$  GHz. The  $IL = 2.4$  dB is 0.8 dB higher than simulated. This is characteristic of additional radiative and Ohmic losses.

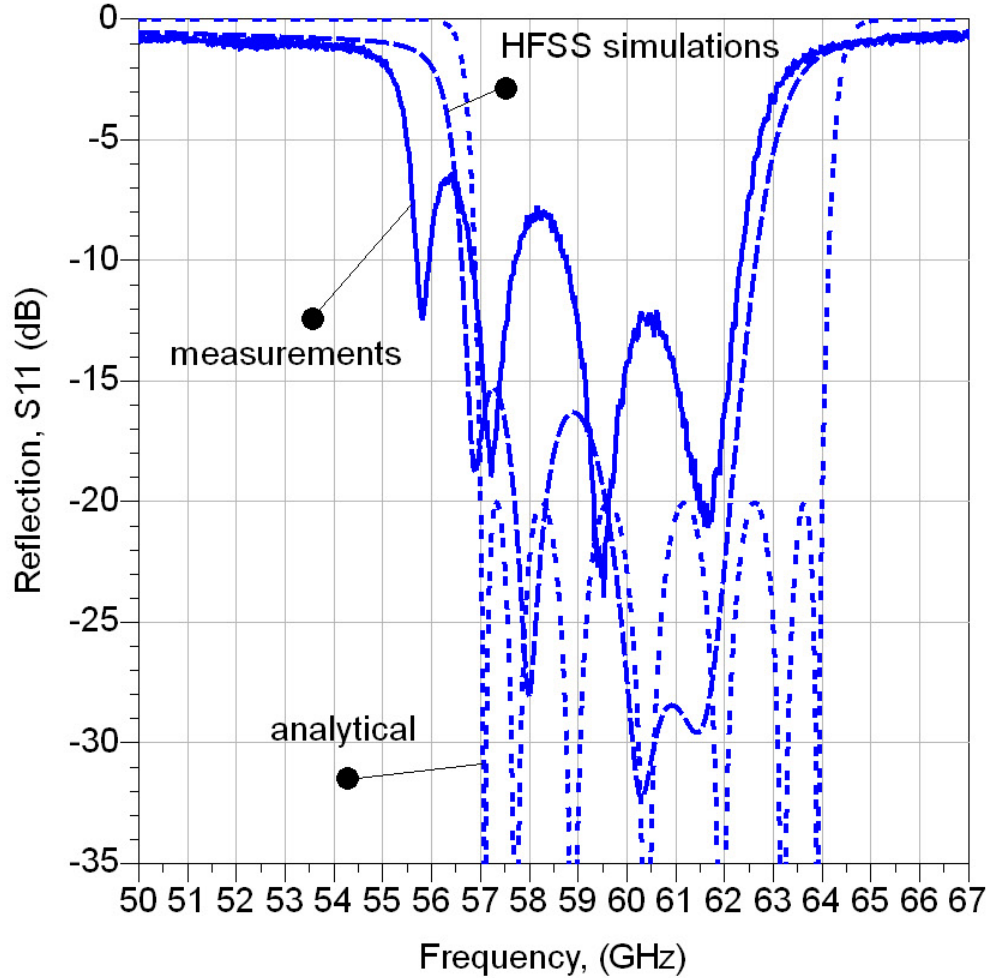


Fig. 4.8.: Comparison between measurements and simulations of the reflection parameter  $S_{11}$ .

Concomitantly with a larger bandwidth, a higher return-loss in the passband is measured and the filter is shifted towards lower frequencies i.e. 1 GHz offset from  $f_c$ . This is a characteristic of larger physical dimensions which may be resulting from process variations. Additionally, this may also be affecting the irises openings resulting in a higher than expected return-loss in the lower part of the passband. In the stop band below the passband the  $IL$  is measured about 50 dB and  $IL = 30$  dB above the passband. The attenuation below the passband is attributed to the excitation of spurious surface waves and partially due to the measurement device limitations.

The lower attenuation above the passband, is attributed to the excitation of higher modes. Alternative design solutions suggest the introduction of transmission zeros, by using additional negative coupling between resonators [Lee 09, Randall 01]. This is in increasing further the selectivity and symmetry of the band pass filter.

Table 4.2.: Filters' parameters designed and optimized with HFSS ( $N = 7$ , Chebyshev filter 0.0432 dB equal ripple).

|             | Design      | Optim.      | $\Delta$    |
|-------------|-------------|-------------|-------------|
|             | ( $\mu m$ ) | ( $\mu m$ ) | ( $\mu m$ ) |
| $d_1 = d_6$ | 900         | 894         | -6          |
| $d_2 = d_5$ | 800         | 628         | -172        |
| $d_3 = d_4$ | 750         | 587         | -163        |
| $l_0 = l_6$ | 1442        | 1442        | 0           |
| $l_1$       | 1066        | 1108        | 42          |
| $l_2$       | 1216        | 1299        | 83          |
| $l_3$       | 1257        | 1318        | 61          |
| $l_4$       | 1262        | 1299        | 37          |
| $l_5$       | 1257        | 1108        | -149        |

#### 4.4 CONCLUSIONS

A Chebyshev seven-pole filter, with a fractional bandwidth,  $FBW = 11.8\%$  ( $BW = 7$  from 55.6 to 62.6 GHz) around a central frequency,  $f_c = 59.1$  GHz was demonstrated. The measurements show a low passband insertion loss ( $IL = 2.4$  dB) and a high stop band below ( $IL = 50$  dB) and above ( $IL = 30$  dB) the band pass. The filter is a substrate integrated waveguide (SIW) in LTCC, with dimensions  $9450 \times 2100 \times 200 \mu m^3$ .

The design of the filter is based on a filter synthesis method using well established network methods and tabulated parameters. Those parameters are used to identify the electrical characteristics that are translated into physical dimensions following a model extraction of parameters from finite element simulations. Additionally, a new feeding design was presented to fulfill the large-band design requirements

Compared to the filter of Lee *et al.* [Lee 09], the insertion loss is slightly higher, while the filter is more than two times longer ( $IL = 2.4$  dB vs.  $IL = 1.77$  dB) and the seven composing resonators are characterized by lower quality factors. Therefore, the better insertion loss performance is explained by a better feeding mechanism. Additionally, the filter has a larger bandwidth (11.8% vs. 11%). However, it has a relatively high

return-loss in the pass band and it suffers of a 1 GHz frequency shift. These are imputed to variations of the LTCC process and can be corrected in a second iteration, provided the measurements are in very good agreement with the simulations. This demonstrates that a large band multi-poles filter, with high rejection out of band is feasible at 60 GHz in LTCC.



## CHAPTER 5

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# ACTIVE APPLICATION: ULTRA-LOW PHASE-NOISE OSCILLATORS

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### 5.1 INTRODUCTION

The oscillator is an essential component for microwave systems such as those used in communications, RADARs, instrumentation, navigation and electronic warfare [Robertson 96, Pozar 05, Kajfez 86, Rogers 91]. An overview of solid-state oscillators can be found in [Robertson 96] where early developments based on Gunn and IMPATT diodes are presented as well as after 1970s GaAs MESFET devices operating in the microwave and mm-wave frequencies. More recently, SiGe and III-V HBTs [Wanner 06] as well as HEMTs (FET) [Chen 03b] and CMOS [Cho 05] have broadened the range of possibilities and demonstrators. Different amplifier topologies having their own advantages: the common emitter/source length to ground increases instability and uses base/gate series feedback [Caruth 00, Dancila 11b], common base/gate for wide-band oscillators [Aoki 96, Kirby 06] and common collector/drain for medium power oscillators [Rogers 91]. Other topologies encountered for high frequency oscillators are the push-push oscillators [Baeyens 03] where two symmetrical oscillators operate at half the output frequency in anti-phase i.e.  $180^\circ$  out of phase with each other and are combined to yield a strong second harmonic output signal; balanced and differential Colpitts [Baeyens 00, Lao 04] and hybrid DR based oscillators [Funabashi 94]-[Popovic 97].

The common principle of operation is that an active non-linear device, such as a diode or a transistor is used in conjunction with a passive circuit to convert DC to a sinusoidal steady-state RF signal [Pozar 05]. The transistor and its associated circuitry forms the amplifier that supplies the energy to compensate the losses in the oscillator.

The frequency of operation is limiting the choice of resonators among which are lumped components, transmission lines, dielectric resonators, air cavity resonators, crystal resonators and other YIG, SAW and BAW resonators. Each resonator is characterized by a Q-factor which defines its selectivity. At very high frequencies, lumped and transmission line resonators are limited in performance [Tilmans 03]. The requirement for an excellent oscillator is evaluated in terms of phase-noise. Ultra-low phase-noise oscillator correspond to very pure and stable wave-form generators. The phase-noise is particularly influenced by the loaded Q-factor of the resonator and the output power of the oscillator. Therefore, the oscillator phase-noise is improved by using resonators with a high loaded Q-factor, ( $Q_l$ ) as can be seen from the linear analysis by Robin [Robins 82] of Leeson's equation [Leeson 66], simplified in [Rogers 91]:

$$\mathcal{L}(f_m) = \frac{FkTf_0^2}{8PQ_l^2f_m^2} \quad (5.1)$$

where  $\mathcal{L}(f_m)$  is the phase noise,  $F$  is the active device noise figure at operating power,  $kT$  is the thermal noise (equal to -174 dBm in a 1 Hz bandwidth at room temperature),  $f_{osc}$  is the oscillation frequency,  $P_{OUT}$  is the output power,  $Q_l$  is the loaded Q factor and  $f_m$  is the offset frequency from the carrier, the oscillation frequency.

## 5.2 OPERATION PRINCIPLES

A qualitative description of the oscillator operation principle is provided in [Rogers 91]. The build-up of the oscillating signal is described starting from the white noise of the amplifier, given for illustration in Figure 5.1. The upper curve shows the oscillator's open-loop characteristic which results from a disconnection at the junction between the amplifier and the resonator, as can be seen in Figure 5.2-b). The signal reflects back and forth between the negative resistance and the resonator.  $\Gamma_A$  and  $\Gamma_R$  are respectively the reflection coefficient of the active circuit at its negative impedance port and the reflection coefficient of the resonator. If  $|\Gamma_A\Gamma_R| > 1$  the signal is gaining power at each loop increasing its amplitude above the magnitude of its previous pass i.e. A,B,C,D, etc. in Figure 5.1. The narrowing open loop gain of the oscillator results from the selectivity of the resonator, as only the signal arriving in a constructive phase is further amplified causing the noise in the oscillator to peak up. The part of the signal outside the oscillator's bandwidth is attenuated refining the power density to a single frequency. The result is a signal dynamically manipulated by the electrical circuit in the presence of noise [Rogers 91].

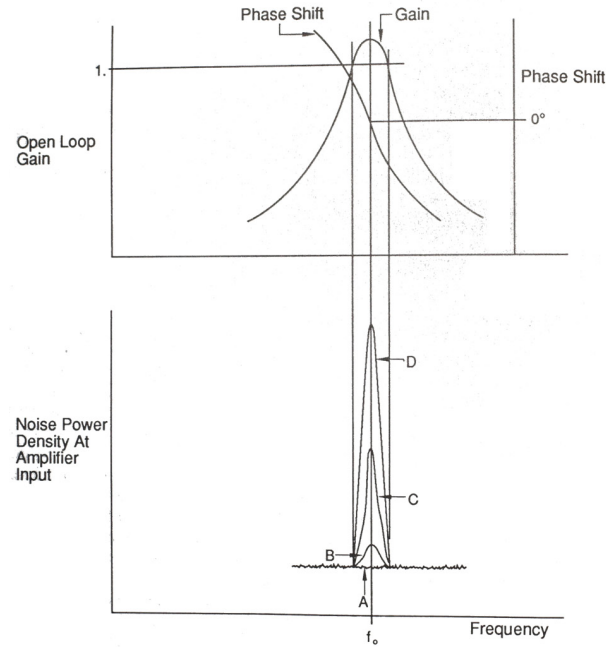


Fig. 5.1.: Oscillation magnitude growth [Rogers 91].

Eventually, the magnitude's growth produces a signal that reaches the level where non linearities appear in the active circuit. This may result in signal compression reducing the loop gain and causing the closed loop gain to move towards a gain close to unity, while increasing the phase noise. Nevertheless, the open loop gain of the amplifier does not need to be unitary but is required to be at an adequate magnitude value (i.e. 4 dB for commercial usage, 8 dB for harsh environment), in order to allow the start-up of the oscillation at the low limit range temperatures [Rogers 91].

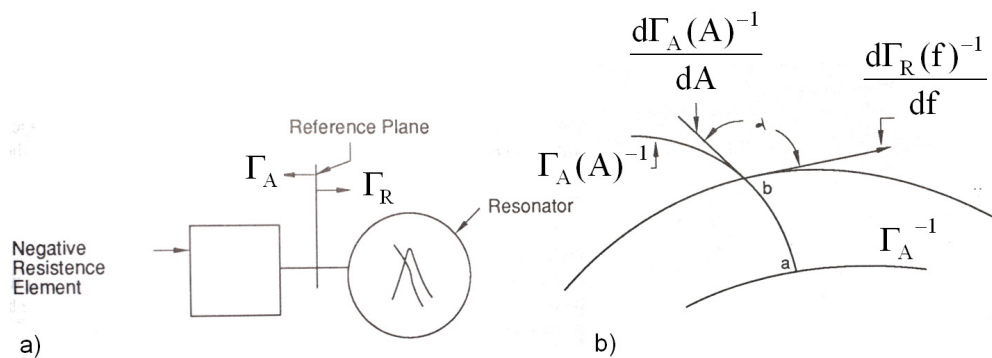


Fig. 5.2.: a) Block diagram of the reflection form of oscillator. b) Reflection coefficient and oscillation condition [Rogers 91].

Accordingly to Rogers analysis' developed in: "*Low phase-noise microwave oscillator design*" [Rogers 91], we consider the reflection oscillator schematically described in Figure 5.2-a).

$\Gamma_A$  and  $\Gamma_R$  are respectively the reflection coefficient of the active circuit, at its negative impedance port and the reflection coefficient of the resonator. They are both frequency dependent and  $\Gamma_A$  is also a function of magnitude, while  $\Gamma_R$  is not. At turn-on, Figure 5.2-b) point (a),  $\Gamma_A$  is considered to have a value which is not influenced by the signal level. After turn-on,  $\Gamma_A(A)^{-1}$  rises along its trajectory towards its steady-state oscillation frequency, Figure 5.2-b) point (b), where  $A$  is the amplitude of the active circuit. The rise of  $\Gamma_A(A)^{-1}$  is meaningful of the amplifier's non linearities, concomitant with a  $\Gamma_A(A)$  decline. At steady-state oscillation, at point (b), the inverse of the active circuit gain has the instantaneous magnitude trajectory vector:  $\frac{d\Gamma_A^{-1}(A)}{dA}$  and  $\Gamma_R$  has the instantaneous frequency trajectory vector:  $\frac{d\Gamma_R^{-1}(f)}{df}$ . These two vectors have an angle  $\alpha$  in between them and for sustained oscillations:  $0 < \alpha < 180^\circ$ . Additionally, phase noise minimization requires [Rogers 91]:

1.  $\alpha = 90^\circ$
2.  $\frac{d\Gamma_A^{-1}(A)}{dA}$  as to be minimized
3.  $\frac{d\Gamma_R^{-1}(f)}{df}$  as to be maximized

Condition (1.) requires the orthogonality between the magnitude rate of change of  $|\Gamma_A|$  and the frequency rate of change of  $\Gamma_R$ . This orthogonality means that the fluctuations of the frequency will not modulate the magnitude and that the magnitude fluctuations will not modulate the phase and therefore the resonance frequency. Condition (2.) requires the maximization of the magnitude stability and condition (3.) requires the maximization of the phase shift with frequency and therefore implies the maximization of the loaded Q-factor,  $Q_l$ .

### 5.3 60 GHZ INTEGRATED OSCILLATOR IN MCM-D

Integrated passives components, as LC tanks in monolithic microwave integrated circuits (MMIC) are limited in performance at mm-wave and degrade oscillators' performance [Tilmans 03]. In order to maximize  $Q_l$  and therefore to reduce the phase noise according to (5.1), air cavity resonators as well as dielectric resonators (DR) were assembled in hybrid configurations as they provide very high Q-factors [Kajfez 86]. However, the accurate placement of DR is cost prohibitive, especially at high frequencies. As a lower cost alternative, high-Q air cavities were realized using micromachining techniques [Harle 04b]-[Ocket 05]. Micromachined cavities show similar characteristics to the metallic based air cavity resonators while being fully integrated. Those resonators were effectively used to stabilize Ka-band [Kwon 99] and V-band [Song 09] oscillators. However, complex additional processing is required to bond the cavities and

manufacturing challenges make their integration more difficult. Furthermore, such hybrid oscillators are relatively large in size in comparison with full MMIC implemented solutions.

Similarly, cavity resonators were integrated in LTCC [Lee 05a] and no extra bonding for the cavity is required nor other additional operations than the ones required by the basic fabrication [VTT 08]. However, the performance is lower than for air cavity resonators, as dielectric losses have to be taken into account. Nevertheless, the lateral dimensions of the LTCC integrated cavity resonators are reduced by a factor  $\sqrt{\epsilon_r} = \sqrt{5.7} = 2.4$  as compared to air cavities. Recently, cavity resonator filters realized in MCM-D technology demonstrated very good performance in terms of insertion loss and Q-factors [Posada 08]. In this approach, the lateral walls of the cavity resonator were defined by two rows of TSVs, with a diameter of  $100 \mu\text{m}$  and a pitch of  $200 \mu\text{m}$ .

In this context, an active application realized within this PhD, consists in V-band low-phase-noise oscillators, using the SiCR at 60 GHz, fully integrated in IMEC's MCM-D technology, presented in Section 2.5 and called Silicon integrated Cavity Resonator based Oscillator, SiCRO. The SiCRO is presented in Figure 5.3.

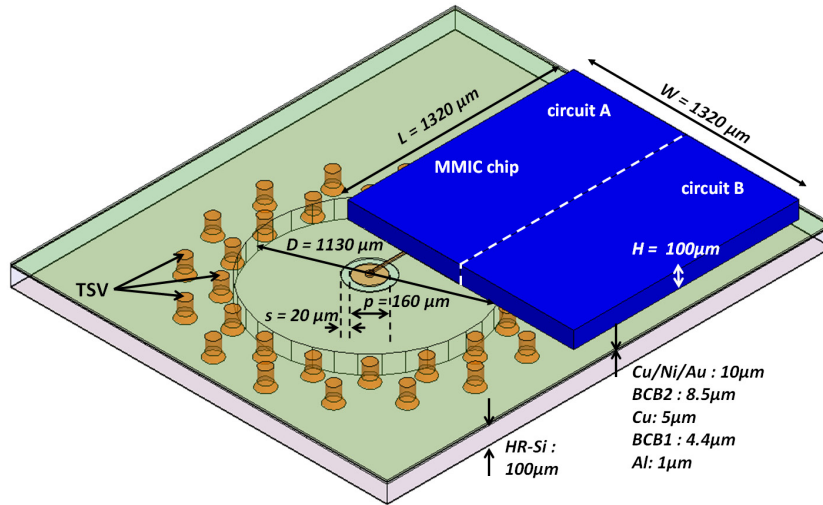


Fig. 5.3.: V-band Substrate Integrated Cavity Resonator based Oscillator, SiCRO.

Different resonance frequencies and different levels of coupling are demonstrated. For this purpose a new feeding structure is proposed. For the design of the oscillators, several analysis methods are complementary used. Measurements demonstrate the improvement of the phase-noise characteristics of the oscillators using silicon integrated cavity resonators (SiCRs) over the oscillators using shorted CPW lines as resonators and using the same active device. In comparison with the air cavity resonators, the side dimensions of the SiCRs are reduced in size by a factor equal to  $\sqrt{\epsilon_r} = \sqrt{11.9} = 3.4$ , which represents a promising premise in comparison with air cavity resonators. Additionally, the surface taken by the resonators integrated in the substrate is partially

used as landing structure for the active MMIC. This architecture further improves the level of integration and contributes substantially to the cost reduction while preserving very high figures of merit. An *a-posteriori* analysis assesses the hybrid implementation and proposes future improvements.

### 5.3.1 Silicon Integrated Cavity Resonator

The SiCR is implemented in the technology described in Section 2.5.1. Additionally, integrated cavity resonators and MMIC die, flip-chip assembled on top of the MCM-D are presented in Figure 5.4 which shows the cross section of the MCM-D integration platform as described in [Posada 07]. The integrated passives are realized in the top thin film layers formed by three metallizations embedded in benzo-cyclobutene (BCB)  $\varepsilon_{BCB} = 2.65$ . The first layer,  $1\text{ }\mu\text{m}$  Al is sputtered directly on top of the HR-Si ( $\varepsilon_{HR-Si} = 11.9, \tan \delta = 0.0026$ ) and covered by  $4.4\text{ }\mu\text{m}$  thick BCB. The second layer,  $5\text{ }\mu\text{m}$  Cu is covered by  $8.5\text{ }\mu\text{m}$  BCB. The third layer is an electroplated stack of  $10\text{ }\mu\text{m}$  Cu/Ni/Au. TSVs are dry etched through the  $100\text{ }\mu\text{m}$  thinned HR-Si wafer and conformally plated with  $5\text{ }\mu\text{m}$  Cu. The  $5\text{ }\mu\text{m}$  electroplated Cu also covers the back side of the wafer. The TSV have a bottom diameter of  $100\text{ }\mu\text{m}$  and a top diameter of  $75\text{ }\mu\text{m}$ . They are placed with a pitch of  $200\text{ }\mu\text{m}$ , imposed by the technology.

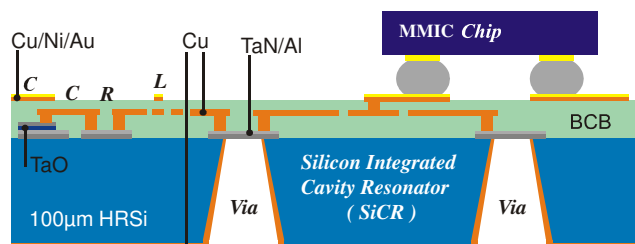


Fig. 5.4.: Cross section of the thin-film MCM-D, 100  $\mu\text{m}$  thinned HR-Si substrate with SiCR shaped by TSVs. The MMIC chip is directly flip-chipped on landing structures patterned in the third metallization.

The cavity resonator used is a cylindrical resonator in its  $\text{TM}_{010}$  mode of resonance, filled with a material of relative permittivity,  $\epsilon_r$  and permeability  $\mu_r = 1$ , as presented and characterized in Section 2.5.2. We recall here only that considering a cylindrical cavity flanked by a double row of TSVs, with a particular shape described in Figure 2.22, (bottom diameter of 100  $\mu\text{m}$  and top diameter of 75  $\mu\text{m}$ ), a diameter  $D = 1130 \mu\text{m}$  allows  $f_0 = 60 \text{ GHz}$ , as the effective diameter of the TSV's shielding is larger than the diameter used in analytical and eigen mode simulations. In this case, TSVs form the cavity resonator with the diameter  $D$  taken as the diameter of the cylinder formed by the base of the TSVs, as can be seen in Figure 5.5. A double row of TSVs is used, 200  $\mu\text{m}$  pitch, much less than half wavelength which is approximately the length of the diameter  $D$ . This configuration demonstrated very good shielding performance at 60 GHz, following [Posada 08].

### 5.3.2 Coupling design

The patch feeding, investigated in Section 2.5 is adapted to set the reflection coefficient for oscillator design purposes. The new coupling structure is composed of a metallic patch realized in the MCM-D's second metallization connected by microstrip lines realized in the top third metallization, visible in Figure 5.5. The MMIC is directly flip-chip connected to the probing pads of the third metallization. The coupling is controlled by the diameter,  $p$  of the patch fed, which is precisely controlled by the thin film patterning techniques.

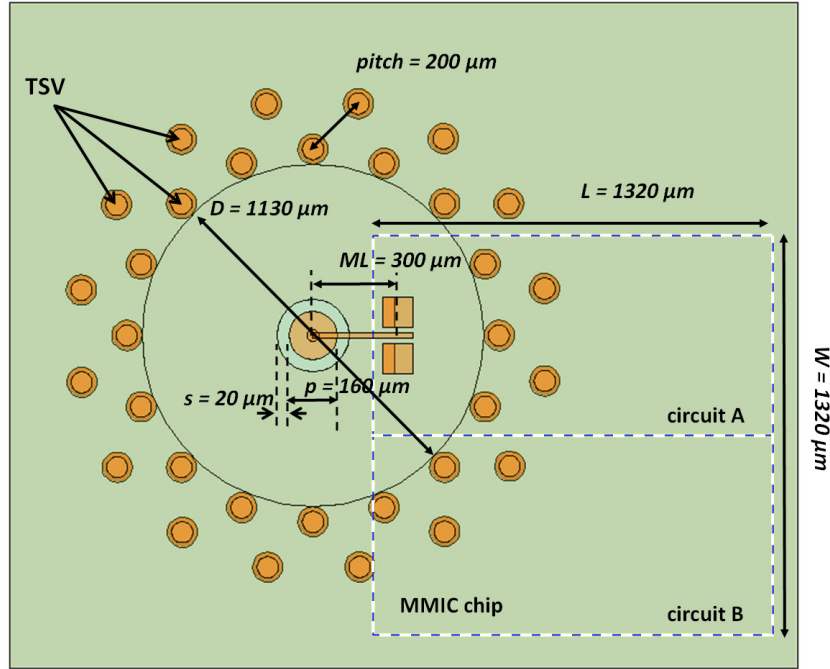


Fig. 5.5.: Top view of the SiCRO. The TSVs define the cylindrical cavity resonator. The patch feed is located in metal 2. The microstrip line and measurement pads are in metal 3. The diameter of the SiCR,  $D = 1130 \mu m$ , the diameter of the feeding patch,  $p = 160 \mu m$  and the slot,  $s = 20 \mu m$ .

The patch feed in Figure 5.5, previously introduced in Figure 2.20 and Figure 2.21, couples electrically to the resonator and excites the  $TM_{010}$  fundamental mode of the cylindrical cavity resonator, as detailed in Figure 2.29. This configuration allows the normalized impedance  $z$  at resonance and the coupling coefficient  $\kappa$  to be chosen by changing the diameter of the of the coupling patch,  $p$  defined in Figure 2.20 and Figure 5.5.

### 5.3.3 Oscillator design

In this section, a low-phase noise oscillator at 60 GHz is designed *a-priori*. Different analysis using small signal simulations describe complementary the oscillator's behavior and are compared to harmonic balance simulations (HB), performed with the software Advanced Design System (ADS) [ADS]. HB is using large signal models and allows to simulate the output power of the oscillator.

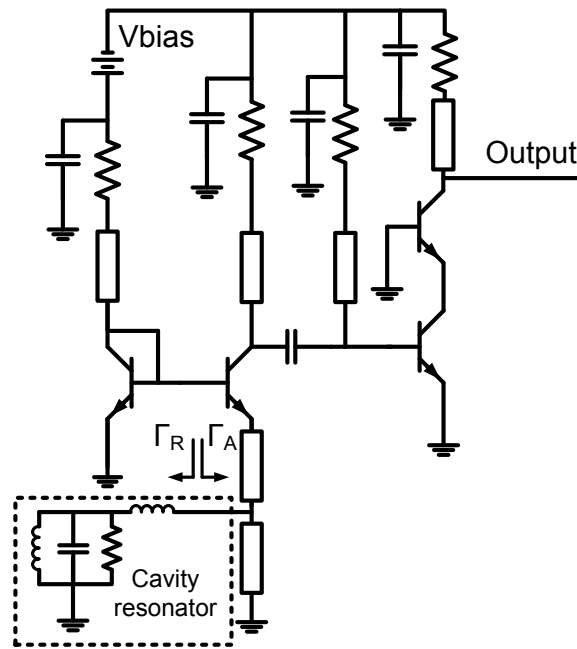


Fig. 5.6.: Oscillator schematic in a common emitter HBT configuration, with a mirror of current, cavity resonator model and single stage cascode buffer. Designed by Andreas John from AC Microwave GmbH, now with Hella KGaA Hueck & Co.

The circuit diagram of the SiCRO is presented in Figure 5.6. The circuit was initially designed for an air cavity resonator, as described in [Ocket 08]. In the following analysis, the active device presented in Figure 5.7 is used in combination with a SiCR. The MMIC is flip-chip connected on the MCM-D with the emitter of the HBT connected to the SiCR, one port device. The oscillator is designed in a series feedback scheme with the output taken from the collector and further amplified by an on-chip single stage cascode buffer. Biasing lines are also present in the circuitry, as can be seen in the micrograph of the active device (Figure 5.7). In the MMIC die ( $1320 \times 1320 \mu\text{m}$ ) two active circuits are located: the bottom one, version B is similar after a rotation by  $180^\circ$  to the top one, version A. The two versions are different by a small variation in the on chip length of the inductive strip at the base of the transistor. In version A, the connection with the off-chip resonator is on the left hand side and the output is



at the right hand side, as can be seen Figure 5.7. DC-supply is on the top of version A and bottom of version B.

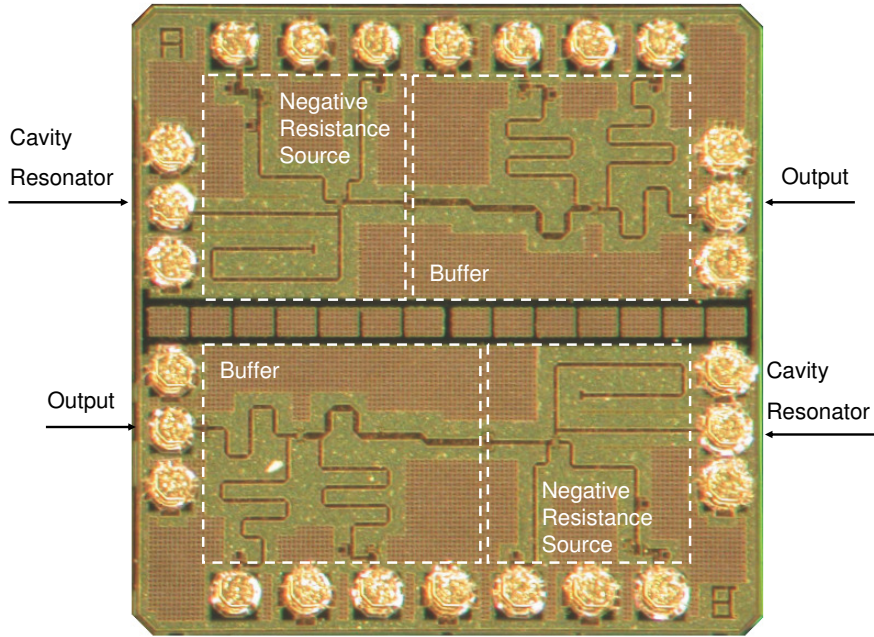


Fig. 5.7.: MMIC die micrograph: two versions available per die. Designed by Andreas John from AC Microwave GmbH, now with Hella KGaA Hueck & Co., realized by NXP Semiconductors in QUBIC4X technology [Deixler 04].

All devices are designed for 1.8 V bias voltage. The biasing current mirror is implemented with the same transistor as the RF gain functional transistor. The design/simulation results for magnitude and phase using the MEXTRAM 504 model for HBT [Deixler 04] are compared with the measurements in Figure 5.8. At room temperature, the magnitude should be approximately 3.5 dB at 60 GHz. While the shapes of the curves are in agreement, the measurements significantly differ from the design: 2.5 dB in magnitude and  $50^\circ$  phase offsets are measured at 60 GHz. This is attributed to the insufficient device models in the higher mm-wave range at the time of design and to the fact that the MMIC was designed considering air above the circuits while when flip-chipped, the transmission lines are landing on the BCB and HR-Si of the MCM-D. As a consequence, the phase shift are more pronounced following the embedding in a higher permittivity than air, as can be seen Figure 5.8.

To de-embed the measurements in the reference plane of the connection with the resonator, a phase shift of  $38^\circ$  is subtracted from the two-port measurements. This phase shift corresponds to the length of the probing CPW line, of  $400\ \mu\text{m}$  long at the end of which the negative resistance port is measured.

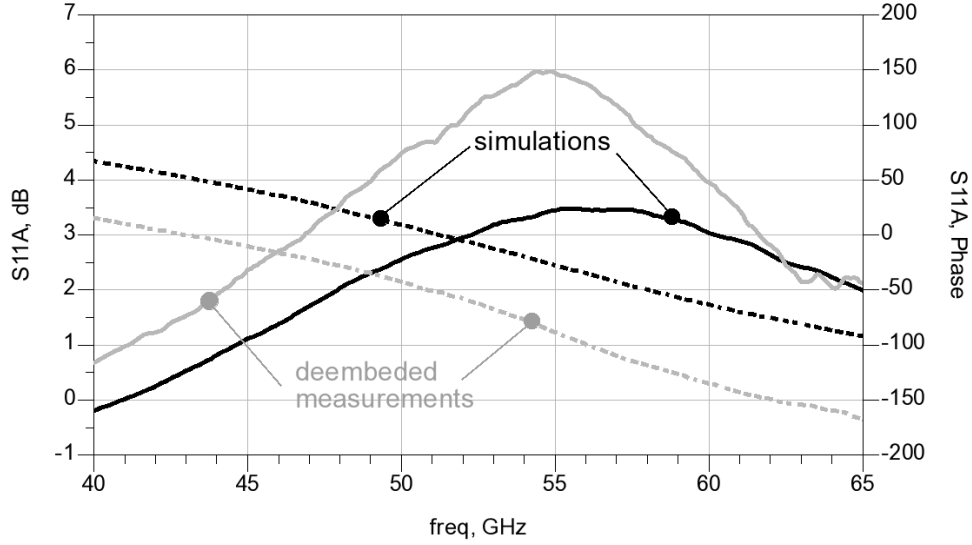


Fig. 5.8.: Comparison between the simulations and the measurements of the magnitude of  $\Gamma_A$ , as defined in Figure 5.6.

The analysis of the oscillation combines in ADS full wave simulations performed with HFSS, on chip interconnections measurements realized directly on the MMIC die, and the MEXTRAM 504 model for HBT [Deixler 04]. The MMIC interconnections were measured separately on chip. The interconnects on the MCM-D as well as the flip-chip solder balls and the SiCR were simulated using HFSS.

Accordingly to Rogers [Rogers 91], the choice was made to consider only overcoupled resonators to be used for the oscillator's reference in conjunction with the provided amplifier. Instead a lightly coupled reflection coefficient would have been chosen, additional oscillation conditions were unveiled which could have turned into spurious oscillations at unwanted frequencies. Of course by doing so, the loaded Q-factor is decreased, as a consequence of the large coupling. So the design of the oscillator was based on trading off the loaded Q-factor,  $Q_l$  in order to fulfill the oscillation condition in terms of magnitude and phase. As the negative resistance provided by the given amplifier is relatively low (i.e.  $S_{11} = 4$  dB at 60 GHz, see Figure 5.8) the resonators used in the oscillator scheme should be highly overcoupled, with coupling coefficients,  $\kappa > 4$ .

To evaluate the frequency of oscillation, the classical method uses the negative impedance concept and evaluates the impedances of the active port ( $Z_A$ ) and the impedance of the resonator port ( $Z_R$ ). The magnitude of  $Z_A$  is negative in the frequency range where there is gain. The oscillation frequency is identified as the frequency where the oscillation conditions are fulfilled:  $Re(Z_R + Z_A) < 0$  and  $Im(Z_R + Z_A) = 0$ . As can be seen in Figure 5.9, the oscillation condition is fulfilled at  $f_{osc} = 60$  GHz.

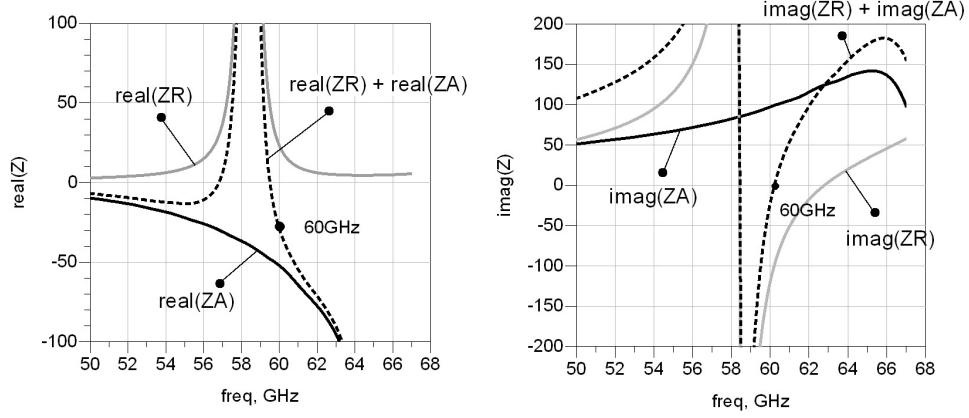


Fig. 5.9.: Oscillation conditions fulfilled at 60 GHz for both the real and imaginary part of the impedances of  $\Gamma_A$ ,  $\Gamma_R$ , as defined in Figure 5.6.

Similarly, the oscillation condition can be expressed in terms of the active device and resonator reflection coefficients, respectively  $S_{11A}$  and  $S_{11R}$ , by considering the complex product  $S_{11R} \cdot S_{11A}$ . Those conditions are visually presented in Figure 5.10. The oscillation condition identifies the frequency of oscillation as the frequency at which the complex product  $S_{11R} \cdot S_{11A}$  crosses the real axes with a magnitude higher than 1. Additionally, the oscillation frequency is identified as the frequency at which both  $S_{11R}$  and  $1/S_{11A}$  are aligned to fulfill the condition on the phase while the condition on the magnitude  $1/S_{11A} < S_{11R}$  is fulfilled [Kajfez 94].

Harada takes into account the exact closed loop gain ( $G$ ) and differs from the previous methods in the bilateral case ( $S_{21} \neq 0$ ) [Harada 99]. The closed loop gain,  $G$  is evaluated by opening the loop and considering a virtual common reference at both ports of the circuit. The frequency of oscillation is identified as the frequency where  $|G| > 1$  and  $\angle G = 0$ , with  $G$  evaluated as follows:

$$G = \frac{S_{21}}{1 - S_{11}S_{22}} \quad (5.2)$$

Finally, Harmonic Balance (HB) simulations are performed using the architecture implemented in ADS. The HB simulation is less intuitive than the previous methods, as the trend of the oscillation frequency as a function of the interconnection lengths is not clearly visible. Nevertheless, the HB analysis enables the identification of the oscillation frequency, phase noise and output power, based on the closed loop gain computation and using the large signal model of the transistor. Therefore this is the most accurate method.

Performing the HB analysis on the 60 GHz SiCRO, results in a phase noise of -108 dBc/Hz @ 1 MHz frequency offset from the carrier frequency. The output power is evaluated to be 1.2 dBm. For comparison, the phase noise is analytically computed using Leeson's equation(5.1) to be -117 dBc/Hz @ 1 MHz frequency offset from the

carrier frequency. This evaluation assumes the active device noise figure at operating power,  $F = 4$  is known for the QUBIC4X technology, the oscillation frequency  $f_{osc} = 60$  GHz, the output power  $P_{out} = 2$  dBm,  $Q_l = 28$  and the offset frequency  $f_m = 1$  MHz.

The resonator used in this particular design, simulated with HFSS, shows an extracted Q-factor  $Q_0 = 124$  and  $Q_l = 13.8$ . Overcoupled resonators are used in the oscillator scheme. The reduction of the loaded Q-factor is a consequence of the high coupling required by the magnitude oscillation conditions and reflection coefficient of the resonator. The level of reflection is obtained by using a relatively large diameter of the feeding patch,  $p = 175 \mu m$ , defined in Figure 5.5 providing a  $\kappa \sim 8$ , at 60 GHz. This value is intentionally over estimated in order to provide still in worse case (i.e. reduced negative impedance) a sustainable oscillation.

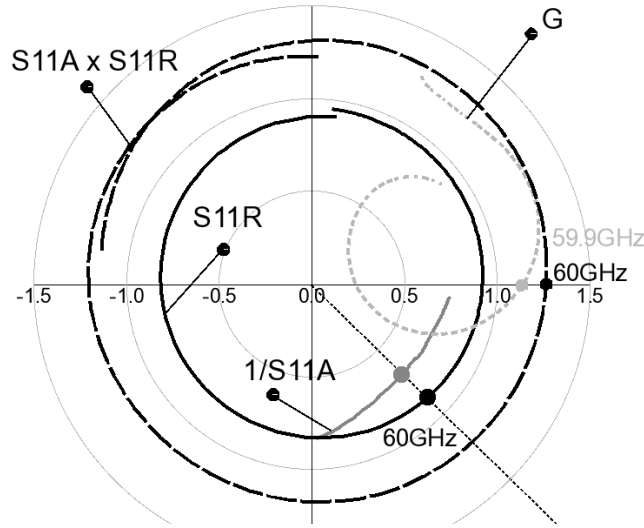


Fig. 5.10.: Graphical representation of the small signal design methods based on the active circuit measurements and cavity resonators simulations with HFSS. Visible are the complex product  $S_{11R} \cdot S_{11A}$ ,  $1/S_{11A}$ ,  $S_{11R}$  and Harada's gain term,  $G$  equation (5.2).

### 5.3.4 Measurements

The final Flip-Chip (FC) mounting of the MMIC on top of the MCM-D substrate with integrated cavity resonator is presented in Figure 5.11. The FC is realized using a FC150 flip-chip tool from Suss Microtec. The MMICs have been provided with gold stud bumps. The MMICs were bonded to gold pads on the MCM-D substrate by means of Au-Au thermo-compression bonding at  $300^\circ C$  for the MMIC and  $100^\circ C$  for the MCM-D substrate. The applied bonding force is dependent on the number of stud

bumps (26 bumps and 300 gF per bump). The dimensions of the assembled oscillator are  $2.6 \times 1.8$  mm and 0.2 mm thick.

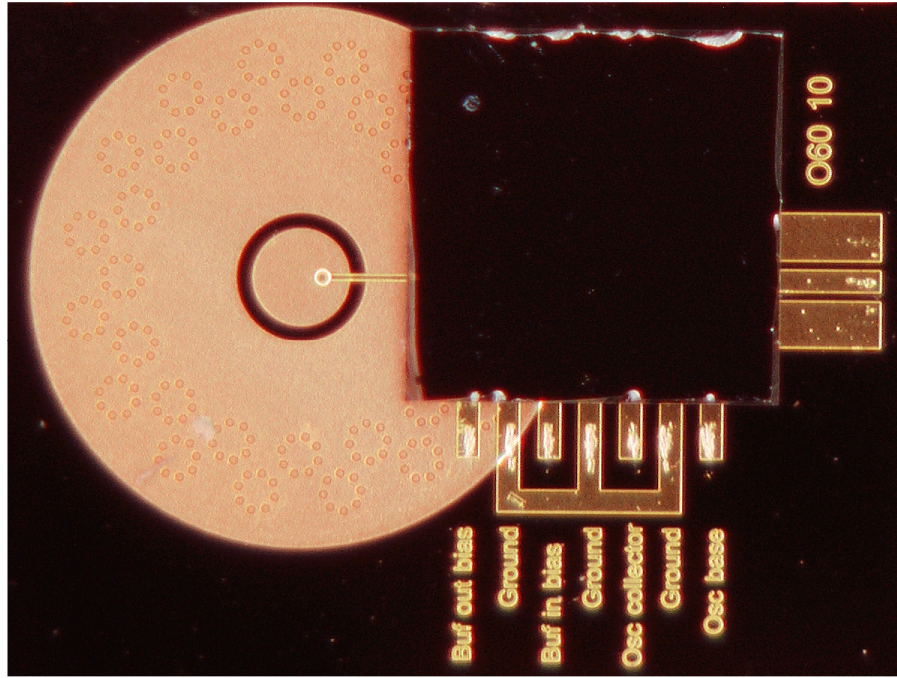


Fig. 5.11.: SiCRO micrograph: MMIC flip-chipped on MCM-D. MMIC die designed by Andreas John from AC Microwave GmbH, now with Hella KGaA Hueck & Co., realized by NXP Semiconductors in QUBIC4X technology [Deixler 04].

The measurements of the oscillators were performed using the FSU 67 GHz spectrum analyzer [FSU]. The oscillator was connected to the Spectrum analyzer by a high frequency cable characterized by an attenuation of 5.5 dB at 60 GHz [Totoku]. The oscillator is probed by 100  $\mu\text{m}$  pitch Ground-Signal-Ground probes characterized by an insertion loss of 1 dB at 60 GHz [Picoprobes]. Therefore, a total attenuation of 6.5 dB has to be considered and is added to the power values measured with the spectrum analyzer presented in Table 5.1.

Under a bias voltage of 1.7 V, the measured oscillation frequency, phase noise and output power are presented in Table 5.1. The phase noise is measured with the Spectrum analyzer using the method described in the application note [Wolf 95]. The spectrum is set as follows: 10 kHz Video Band Width (VBW), 100 kHz Resolution Band Width (RBW), 10 MHz span and the center frequency,  $f_c$  is the carrier frequency or the frequency of oscillation,  $f_{osc}$ . The final measurements are shown in Figure 5.12, where two SiCR oscillators using integrated cavity resonators are compared to an oscillator using a shorted CPW line (820  $\mu\text{m}$  long). The reflection coefficient of the shorted CPW is designed to be similar to the cavity resonator's reflection coefficient but is

somehow higher, resulting in a larger output power. In Table 5.1, the frequency of oscillation,  $f_{osc}$ , the phase noise measured at 1 MHz offset from the carrier frequency, the measured output power are compared for the oscillator using a shorted CPW line as resonator and the SiCRO.

The CPW line resonator's Q-factors are extracted from measurements with the QZERO software, as introduced in Section 2.3.4. At the oscillator's frequency of oscillation  $f_{osc} = 47.3$  GHz, the extracted equivalent Q-factors of the shorted CPW line are:  $Q_0 = 17$  and  $Q_l = 0.9$ .

The two oscillators using SiCR, presented in Figure 5.11, differ by the length of the microstrip line connecting the MMIC to the resonator. The first SiCR oscillator has an interconnection length of  $450\ \mu m$ , while  $580\ \mu m$  for the second.

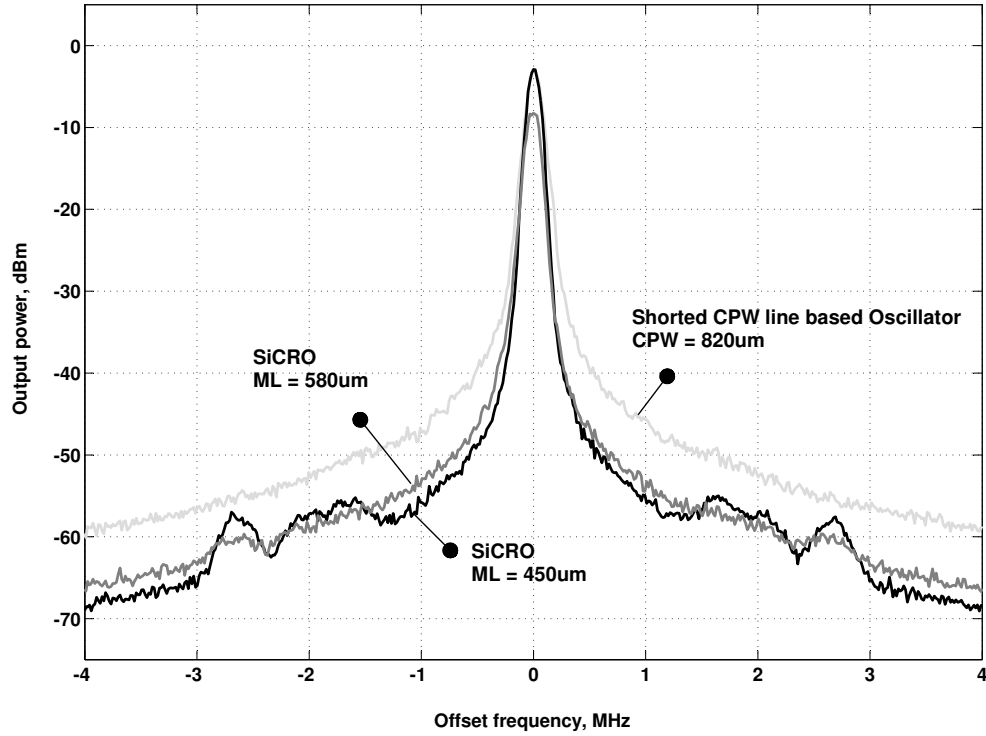


Fig. 5.12.: Comparison between measured oscillator spectrum. The power is defined at the spectrum analyzer port.

The extracted Q-factors at 54 GHz for the SiCR with the  $450\ \mu m$ -long interconnection line, measured prior to FC assembly were:  $Q_0 = 33$ ,  $Q_l = 4.5$ . The improvement in terms of Q-factor is  $Q_{l\ SiCR}/Q_{l\ CPW} = 5$ . The impact on the phase noise is proportional to  $1/Q_l^2$  as following (5.1) and an improvement of 14 dB is expected. This is in good agreement with the measurements as an improvement of 7 dB is demonstrated by SiCROs over the oscillators using a shorted CPW line as resonator, see Table 5.1.



Table 5.1.: Comparison between the proposed SiCROs and the oscillator using a shorted CPW line as resonator.

| Ref.                              | $f_{osc}$<br>(GHz) | Phase Noise<br>(dBc/Hz@1MHz) | $P_{OUT}$<br>(dBm) | $P_{DC}$<br>(mW) |
|-----------------------------------|--------------------|------------------------------|--------------------|------------------|
| This work<br>CPW 820 $\mu m$      | 47.3               | -95                          | 6.8                | 80.3             |
| This work SiCRO<br>ML 580 $\mu m$ | 51.25              | -101                         | 3.5                | 85.0             |
| This work SiCRO<br>ML 450 $\mu m$ | 54                 | -102                         | 5.5                | 90.1             |

### 5.3.5 *A posteriori* analysis

The *a posteriori* analysis consists of analytically combining the reflection coefficients of both the active circuit (measured as a 2-port) and the resonator (measured as a 1-port in reflection, before assembly). On the basis of the oscillation frequency, see Table 5.1, as measured on the fully assembled oscillator, see Figure 5.11, the measurement of the active circuits is de-embedded to bring the analytically combined measurements to fulfill the oscillation conditions at the measured and known frequency of oscillation.

A correction is proposed to bring the oscillation frequency to 60 GHz for the three oscillators measured. The correction consists in reducing the interconnection length in between the active device and resonator. If the length reduction is longer than the initial interconnection length, an additional length of  $\lambda/2$  is further inserted. In this case, the compactness of the solution is decreased and alternatives such as changing the connection location point or the impedance of the connection line could be used to maintain the high level of compactness.

Actually, a phase shift compensation of the interconnection's phase shift by  $-50^\circ$ ,  $-69^\circ$  and  $-80^\circ$  brings the oscillation to 60 GHz for respectively the 54 GHz SiCRO, the 51.25 GHz SiCRO and the 47.3 GHz oscillator with shorted CPW line. Those operations are visually described in Figure 5.13, where the three resonators' reflection measurements are indicated as well as the 60 GHz frequency points. In Figure 5.13, is also indicated the axial line crossing the inverse of the active circuit's reflection factor, at 60 GHz. Reducing the interconnection lines' phase shifts by the indicated values, brings the 60 GHz frequency points into alignment.

Following the phase shift compensations, the 7 dB improvement shown by SiCROs over the oscillator using a shorted CPW line as resonator, presented in Figure 5.12 and Table 5.1 should have been higher. The degradation of the phase noise results from the operation of the oscillators away from their resonance frequency i.e. characterized by a lower loaded Q-factor. However, we consider it as a normal behavior. If they would be operated closer to their unloaded resonance frequency, surely a higher Q-factor ( $Q_0 = 82$ ,  $Q_l = 23$ ) could have been encountered. The phase noise was recalculated

for this case, and shows an improvement of 28 dB (proportional to  $1/Q_l^2$ ) over the oscillator using a shorted CPW line as resonator. In this case, the phase noise is evaluated to be -115 dBc/Hz @ 1 MHz offset from the carrier frequency.

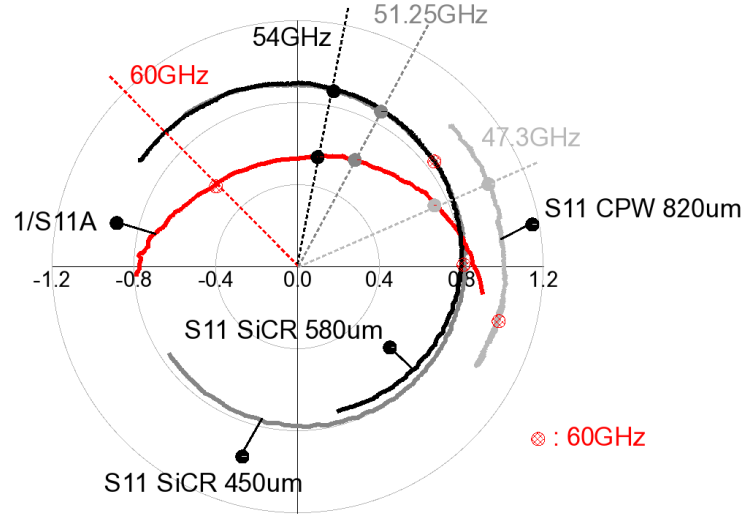


Fig. 5.13.: A posteriori combination of measured reflection factors of active and resonators. Visible are the oscillation conditions fulfilled for the 3 oscillators proposed, presented Table 5.1 and the indication of the 60 GHz on each curve.

#### 5.4 60 GHZ INTEGRATED OSCILLATOR IN LTCC

As a better performance was expected, we continued our research and finally we realized and demonstrated oscillator's with outstanding performance. This performance leaves us no doubts about the opportunities of this hybrid architecture.

For this purpose, a different technology than the MCM-D is used as the fabrication of the MCM-D platform was stopped at IMEC, late 2009. The new oscillator, presented in Figure 5.14 is based on a cavity resonator integrated in LTCC. The resonators integrated in LTCC have similar characteristics than the ones in MCM-D, slightly better. The characteristic unloaded quality is on average  $Q_0 \sim 180$  according to the measurements in LTCC, which should be compared to the characteristics of the resonators in HR-Si of the MCM-D, on average  $Q_0 \sim 120$ . This is due to the height of the LTCC resonator which is nearly double that of the HR-Si resonator ( $193 \mu\text{m}$  vs.  $100 \mu\text{m}$ ) while the electric characteristics are similar. The same active circuit, as previously introduced in Section 5.3 is re-used.



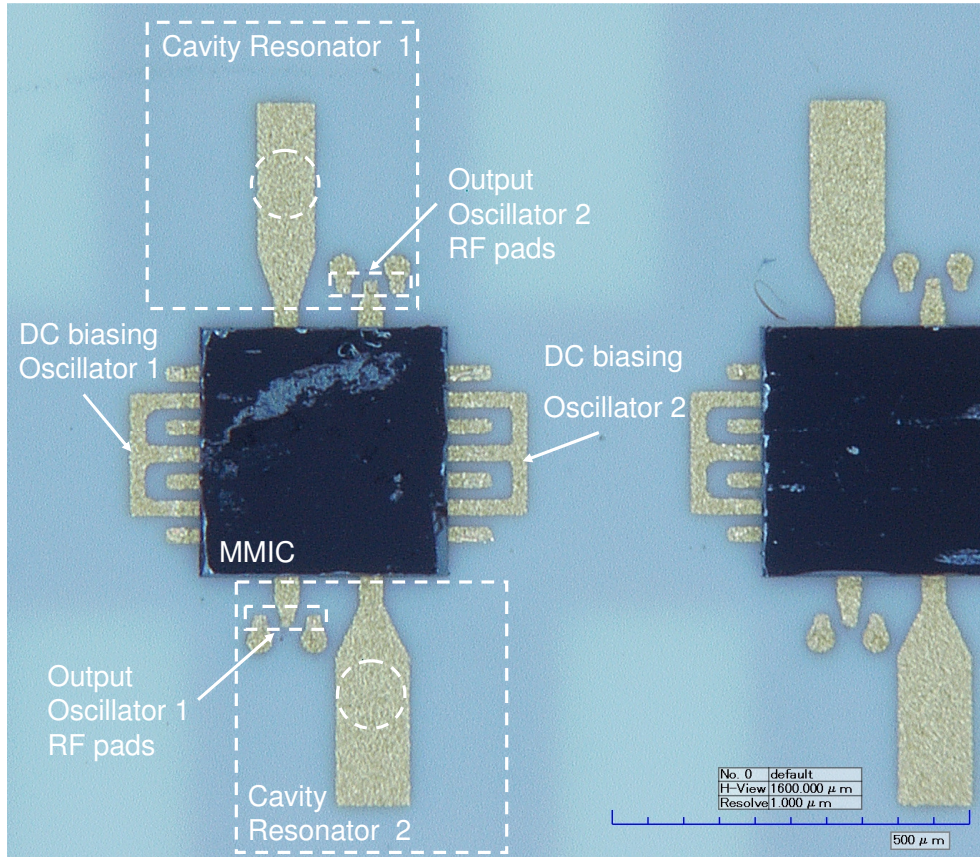


Fig. 5.14.: LTCC SIW cavity resonator based Oscillators, at 60 GHz. Visible are two MMIC dies, incorporating each two active circuits, flip-chip assembled onto a LTCC die integrating SIW cavity resonators. Visible is also the RF connectivity and DC biasing lines. MMIC die designed by Andreas John from AC Microwave GmbH, now with Hella KGaA Hueck & Co., realized by NXP Semiconductors in QUBIC4X technology [Deixler 04].

Firstly, improved performance is presented concerning the frequency of oscillation. We realized and demonstrated an oscillation frequency very close to 60 GHz (the closest:  $f_{osc} = 59.98$  GHz). Secondly, the phase noise is -116 dBc/Hz measured at 1 MHz offset from the carrier, which is very low. This is due to the operation of the cavity resonator characterized by high  $Q$ -factors ( $Q_0 = 123$ ,  $Q_l = 23$ ) at its resonance frequency. To our knowledge, and up to date, the results obtained represents the State of the Art for integrated oscillators operating at 60 GHz as following Section 5.5.

#### 5.4.1 Co-design of the LTCC Integrated Cavity Resonator

Two different cavity resonators are designed in LTCC. The first one is characterized by a height  $b = 193 \mu\text{m}$ , presented in Figure 5.15-a) and the second one is characterized by a height  $b = 695.5 \mu\text{m}$ , presented in Figure 5.15-b).

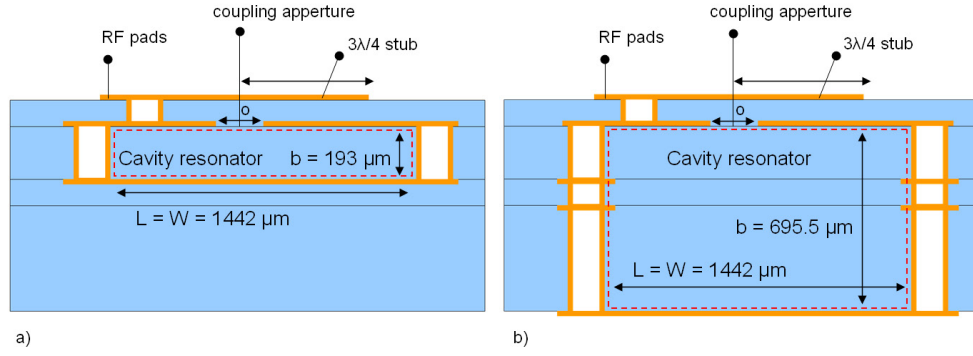


Fig. 5.15.: Cross section of the LTCC cavity resonators. a) Thin cavity resonator of  $193 \mu\text{m}$  and b) thick resonator of  $695.5 \mu\text{m}$ .

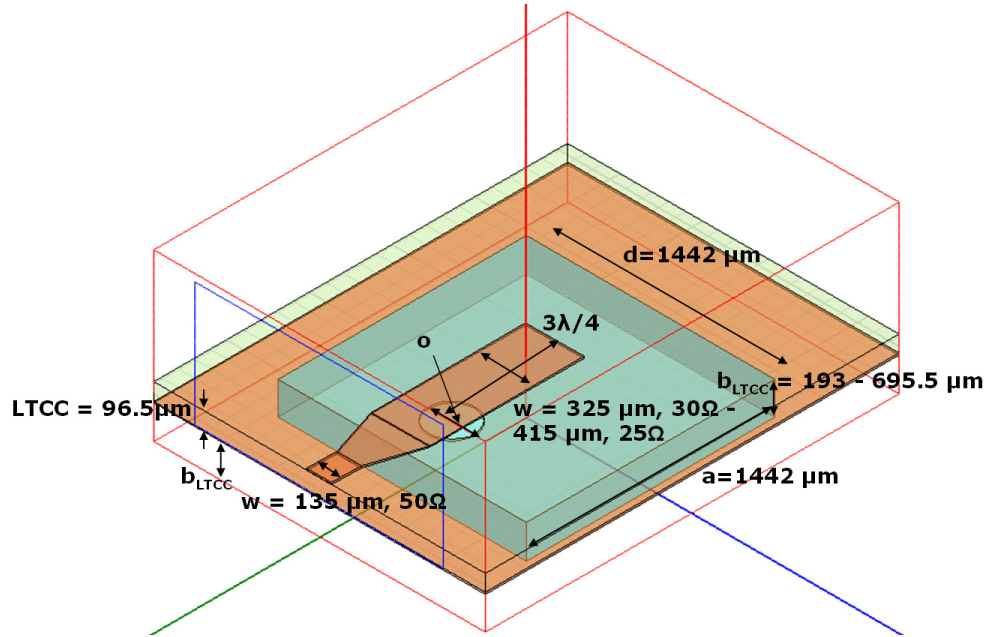


Fig. 5.16.: Parameters definition of LTCC based cavity resonators.

Following eigen mode simulations realized with HFSS, the resonators are characterized by an unloaded  $Q$ -factor of  $Q_0 = 280$  and  $Q_0 = 396$ , respectively with  $b = 193 \mu\text{m}$

and  $b = 695.5 \mu m$ . The Q-factors are extracted with the QZERO software, introduced in Section 2.3.4 [Kajfez 94].

The lateral dimensions of the LTCC cavity resonator are  $a = d = 1442 \mu m$  for a resonance frequency of 60 GHz. A single row of vias (100  $\mu m$  diameter and 200  $\mu m$  pitch) is implemented and the dimensions of the cavity are given tangential to the vias.

A microstrip line coupling structure is provided, similar to the structure analyzed in Section 2.4.2, following Painchaud's seminal work on cavity resonators feeding [Painchaud 76]. The circular aperture of diameter,  $o$ , opened in the ground metallization is related to the width of the microstrip,  $w$  and is taken as main design parameter, fixing the value of the coupling coefficient  $\kappa$ . The configuration and the definition of the parameters are presented in Figure 5.16.

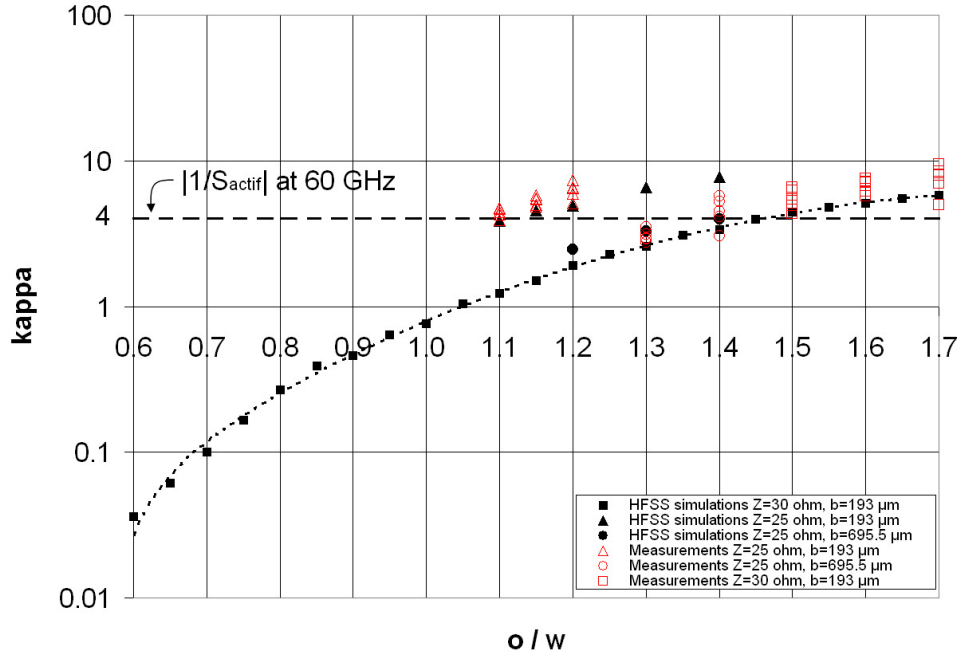


Fig. 5.17.: Coupling coefficient for the microstrip line coupling as a function of the ratio  $o/w$ .

Two different feeding lines with different impedances are considered as following the discussion in Section 2.4.2 related to the coupling of cavity resonators by microstrip lines. It was introduced that to avoid having the aperture diameter much larger than the width of the feeding microstrip line, a solution consists in using lower impedance transmission lines. The impedance are respectively  $Z = 25 \Omega$  with  $w = 325 \mu m$  and  $Z = 30 \Omega$  with  $w = 415 \mu m$ . Actually, lowering the impedance of the line comes with a larger width and lowers the ratio  $o/w$  required for a given coupling coefficient,  $\kappa$ , as can be seen in Figure 5.17. Additionally, the feed with an impedance of  $Z = 25 \Omega$  is

also used to feed the cavities with a height  $b = 695.5 \mu m$ . As a consequence of their larger volume more energy is stored per period and therefore a larger width of the feeding line as well as  $o/w$  ratio is required to obtain the same level of coupling.

As introduced in Section 5.3.3, the oscillation condition related to the magnitude should satisfy  $1/S_{11A} < S_{11R}$ , where  $S_{11A}$  is the negative impedance observed at the active circuit port connected to the resonator and  $S_{11R}$  is the resonator's reflection coefficient [Kajfez 94]. Additionally in Figure 5.17 is plotted the normalized impedance at resonance (60 GHz),  $\beta$ . Considering  $|1/S_{11A}| = 0.6$ , the equivalent impedance is  $200 \Omega$  and  $\kappa = 4$ . For comparison, the measurements of the implementations presented in Figure 5.18 are also presented in Figure 5.17.

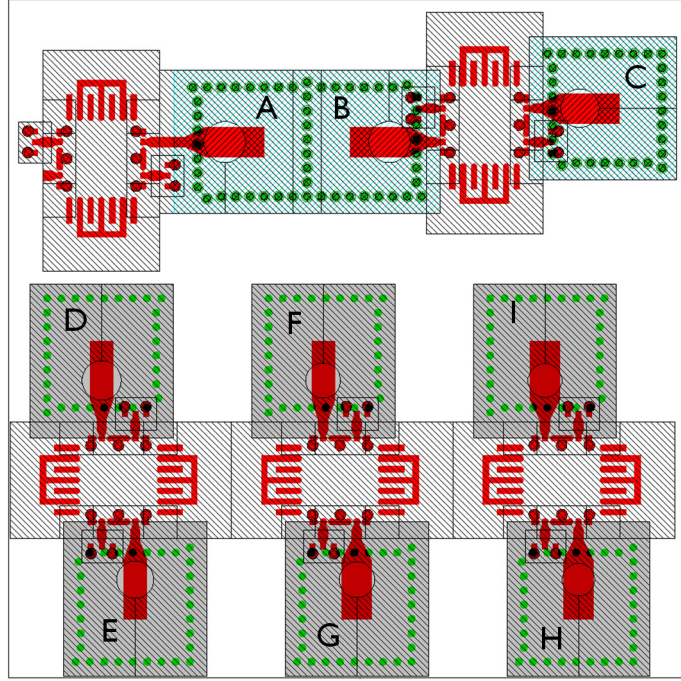


Fig. 5.18.: Layouts of the nine variations implemented. A-B-C are corresponding to  $b = 695.5 \mu m$  and are provided with a feeding line of  $25 \Omega$ . D-E-F are corresponding to  $b = 193 \mu m$  and are provided with a feeding line of  $30 \Omega$ . G-H-I are corresponding to  $b = 193 \mu m$  and are provided with a feeding line of  $25 \Omega$ .

Nine different variations are implemented and are presented in Figure 5.18. Devices A-B-C are corresponding to a height  $b = 695.5 \mu m$  and are provided with a feeding line of  $25 \Omega$ . D-E-F are corresponding to  $b = 193 \mu m$  and are provided with a feeding line of  $25 \Omega$ . G-H-I are corresponding to  $b = 193 \mu m$  and are provided with a feeding line of  $30 \Omega$ .

In total 5 dies, thus 45 resonators were completely characterized and the average results are compared to HFSS simulations and presented in Table 5.2 and Figure 5.17.



Table 5.2.: Comparison between measured and simulated unloaded Q-factor,  $Q_0$ , coupling coefficient,  $\kappa$  and reflection coefficient at 60 GHz, as a function of the ratio  $o/w$ . Considered are resonators with different heights and impedances.

| b           | $Z_c$        | $o/w$ | $Q_0$ |       | $\kappa$ |       | $S_{11}$ @ 60 GHz |                   |
|-------------|--------------|-------|-------|-------|----------|-------|-------------------|-------------------|
| ( $\mu m$ ) | ( $\Omega$ ) |       | Sim.  | Meas. | Sim.     | Meas. | Sim.              | Meas.             |
| 193         | 30           | 1.5   | 272.8 | 124.3 | 4.71     | 5.52  | 0.70 $\angle$ 123 | 0.83 $\angle$ 102 |
| 193         | 30           | 1.6   | 270.6 | 123.8 | 5.69     | 6.74  | 0.75 $\angle$ 121 | 0.79 $\angle$ 102 |
| 193         | 30           | 1.7   | 267.5 | 123.1 | 6.69     | 7.56  | -                 | -                 |
| 193         | 25           | 1.1   | 278.6 | 124.4 | 3.88     | 4.42  | 0.77 $\angle$ 142 | 0.77 $\angle$ 109 |
| 193         | 25           | 1.15  | 276.7 | 123.9 | 4.59     | 5.30  | -                 | -                 |
| 193         | 25           | 1.2   | 271.8 | 122.8 | 4.94     | 6.28  | 0.82 $\angle$ 129 | 0.84 $\angle$ 99  |
| 695.5       | 25           | 1.3   | 396.5 | 279.6 | 3.30     | 3.05  | 0.85 $\angle$ 174 | 0.87 $\angle$ 164 |
| 695.5       | 25           | 1.4   | 392.3 | 278.6 | 3.98     | 4.54  | 0.87 $\angle$ 163 | 0.90 $\angle$ 157 |

A detailed comparison between measured and simulated  $S_{11}$  parameter is presented in Figure 5.19 and Figure 5.20 for the LTCC cavities with  $b = 193 \mu m$  and microstrip line feeding's width respectively  $325 \mu m$ ,  $Z = 30 \Omega$  and  $415 \mu m$ ,  $Z = 25 \Omega$ ; Figure 5.21 for the LTCC cavities with  $b = 695.5 \mu m$  and microstrip line feeding's width of  $415 \mu m$ ,  $Z = 25 \Omega$ .

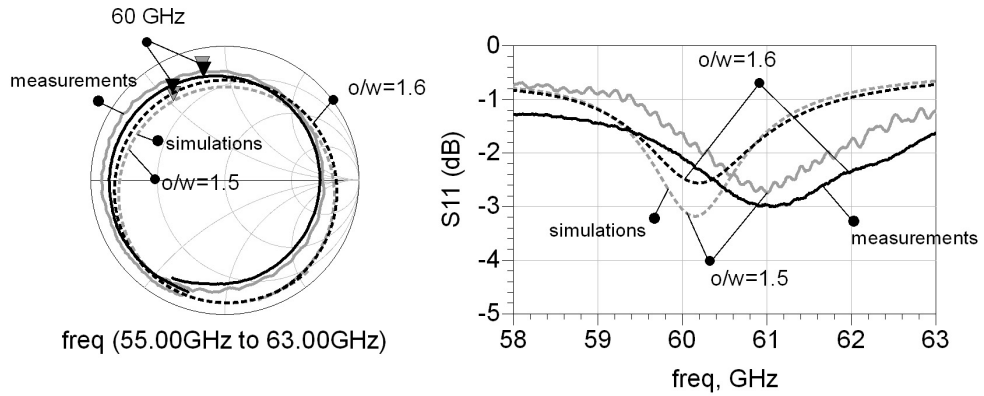


Fig. 5.19.: Simulated and measured  $S_{11}$ , function of  $o/w$  ratio. LTCC cavities of  $b = 193 \mu m$  and microstrip line feeding's width of  $325 \mu m$ ,  $Z = 30 \Omega$ .

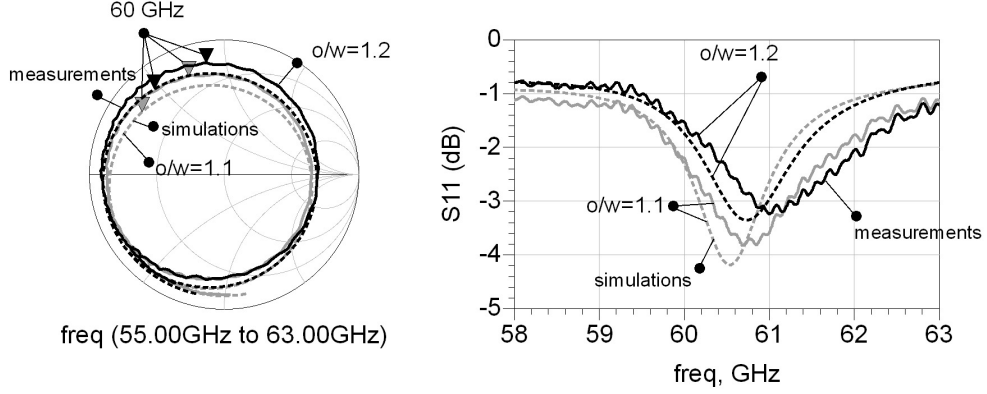


Fig. 5.20.: Simulated and measured  $S_{11}$ , function of  $o/w$  ratio. LTCC cavities of  $b = 193 \mu m$  and microstrip line feeding's width of  $415 \mu m$ ,  $Z = 25 \Omega$ .

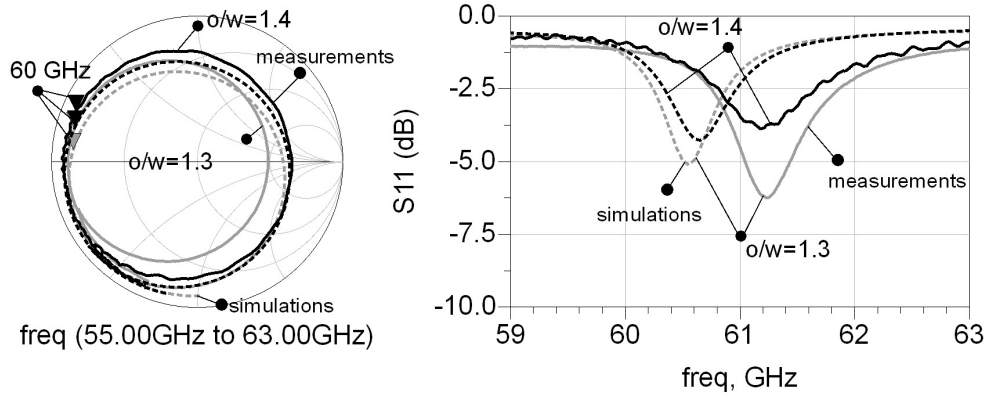


Fig. 5.21.: Simulated and measured  $S_{11}$ , function of  $o/w$  ratio. LTCC cavities of  $b = 695.5 \mu m$  and microstrip line feeding's width of  $415 \mu m$ ,  $Z = 25 \Omega$ .

The coupling coefficient,  $\kappa$  is slightly higher in measurements than in simulations, as can be seen in Table 5.2 and Figure 5.17. This is explained by the fact that the unloaded Q-factor is lower in measurements than in simulations, see Table 5.2, as a lower unloaded Q-factor is reducing the coupling coefficient. This may be due to an over etching of the opening in the metallization. Nevertheless, the values are close and a higher coupling coefficient only increases the chances of success as it decreases the constraints on the magnitude oscillation condition.

The resonance frequency is also slightly different between simulations and measurements. This is particularly the case for larger values of the ratio  $o/w$ . Actually, a larger aperture than the feeding microstrip line may generate additional perturbations in the field distribution inside the cavity resonator by reducing the effective electric width

of the cavity and therefore increasing the resonance frequency as can be seen in Figure 5.19 and Figure 5.21.

The phase difference between simulations and measurements recorded at the 60 GHz frequency point, is on average about  $10^\circ$  for the resonators with  $b = 695.5 \mu m$ ,  $Z = 25 \Omega$ ,  $20^\circ$  and  $30^\circ$  for the resonator with  $b = 193 \mu m$ ,  $Z = 30 \Omega$  and  $Z = 25 \Omega$ , respectively. There is no clear pattern explaining the above mentioned discrepancy. What we could say is that the discrepancies observed concerning the coupling coefficients are of the same order as the discrepancies found for the phase difference. This may suggest that the phase discrepancy is related to the coupling structure, but the phase difference is relatively small and could also be resulting from measurement errors due to the accuracy of the location of the measurement.

#### 5.4.2 Oscillator design

Firstly, an analysis based on a lumped RLC resonator combined to the de-embedded measurements of the active circuit (Figure 5.13) is presented in Figure 5.22.

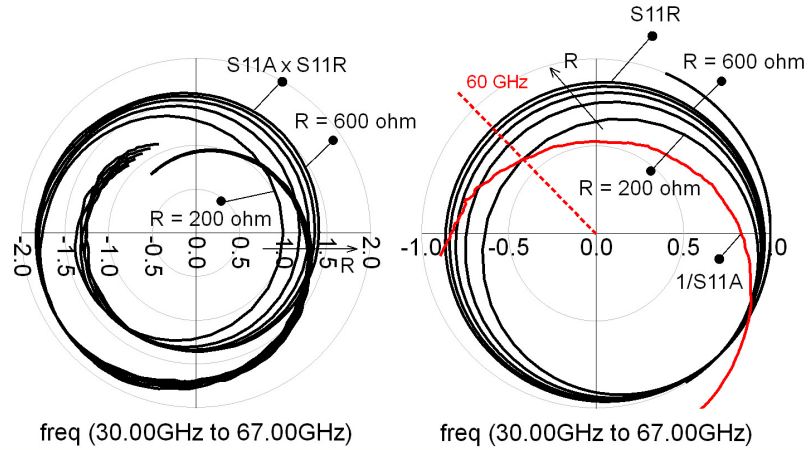


Fig. 5.22.: Analysis based on a lumped RLC resonator combined to the de-embedded measurements of the active circuit presented in Figure 5.13.

An unloaded Q-factor  $Q_0 = 180$  is considered and the lumped elements are obtained as follows:  $C = Q_0 / R 2\pi f_{osc}$  and  $L = 1 / C (2\pi f_{osc})^2$ . An ideal line of  $112^\circ$  is introduced in between the active circuit port and the resonator in order to align the oscillation condition at the expected frequency of oscillation  $f_{osc} = 60$  GHz. The lumped resistance is swept from  $R = 200 \Omega$  to  $R = 600 \Omega$  by steps of  $100 \Omega$  and the results in terms of  $S_{11A}$  and  $S_{11R}$ , considering the complex product  $S_{11R} \cdot S_{11A}$  and the alignment of  $S_{11R}$  and  $1/S_{11A}$  [Kajfez 94] are presented in Figure 5.22. As previously mentioned, a lumped resistance at least equal to  $R = 200 \Omega$ , corresponding to a coupling coefficient  $\kappa = 4$  is required for oscillation.

Secondly, the impact of the phase in between the resonator and active circuit, largely described in Section 5.3.3 directly affects the oscillation frequency. In the RLC case, the impact of  $10^0$  phase shift has an impact of 100 MHz on the frequency of oscillation.

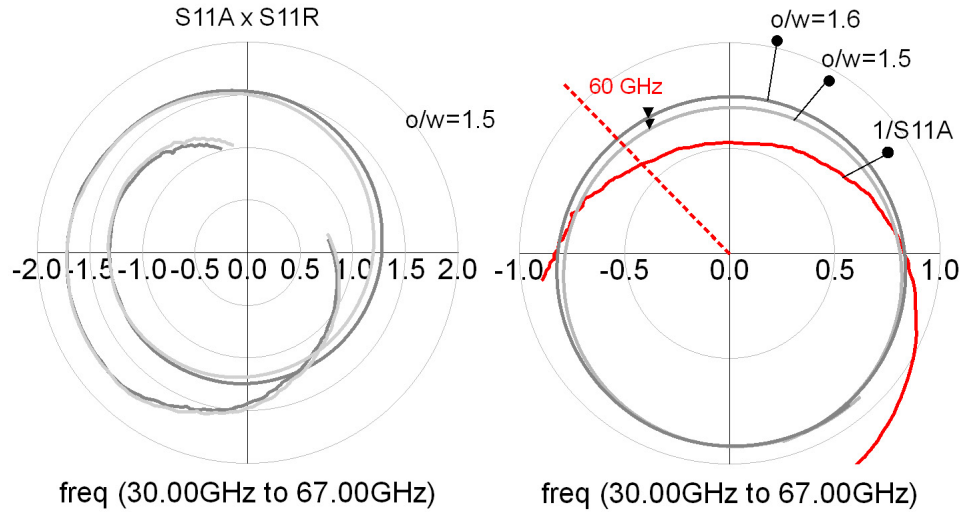


Fig. 5.23.: Oscillation conditions fulfilled around 59.90 GHz for the LTCC cavities,  $b = 193 \mu\text{m}$  and microstrip line feeding's width of  $325 \mu\text{m}$ ,  $Z = 30 \Omega$ .

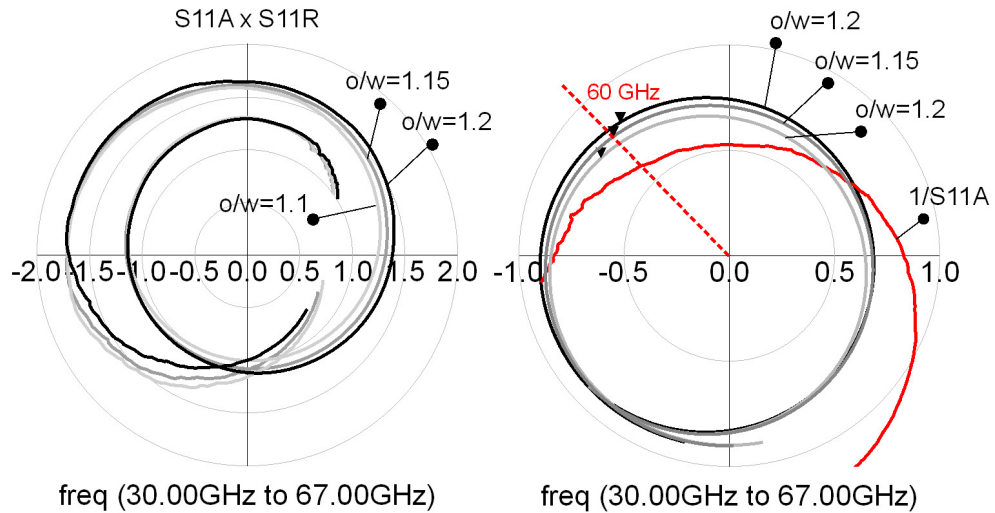


Fig. 5.24.: Oscillation conditions fulfilled around 59.98 GHz for the LTCC cavities,  $b = 193 \mu\text{m}$  and microstrip line feeding's width of  $415 \mu\text{m}$ ,  $Z = 25 \Omega$ .



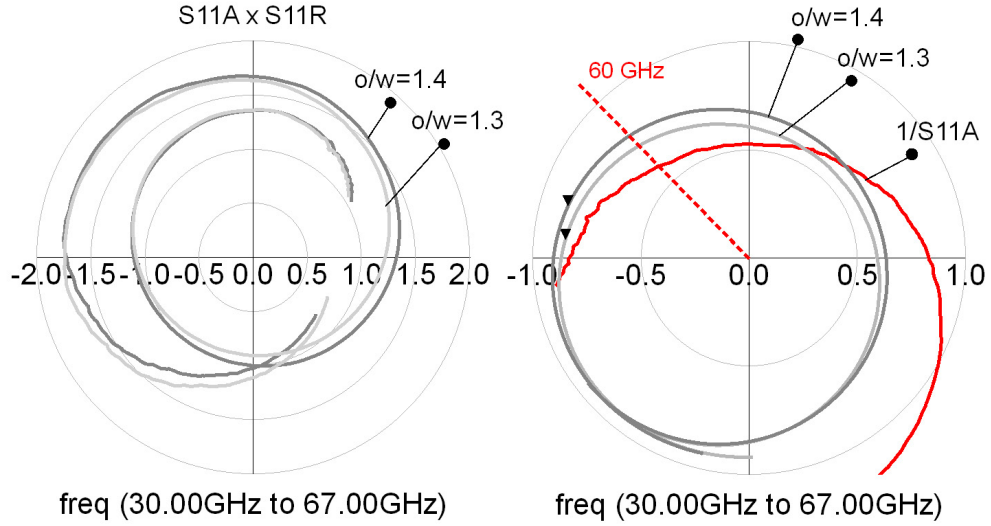


Fig. 5.25.: Oscillation conditions fulfilled around 60.25 GHz for the LTCC cavities,  $b = 695.5 \mu\text{m}$  and microstrip line feeding's width of  $415 \mu\text{m}$ ,  $Z = 25 \Omega$ .

Based on the HFSS simulations of the resonators and de-embedded negative resistance measurement of the amplifier, see Figure 5.8, the small signal analysis is performed. This is described in the polar plots, illustrating the oscillation condition related to  $\Gamma_A$  and  $\Gamma_R$  (defined in Figure 5.6) and applied to three series of oscillators presented in Table 5.3. The first series corresponds to the resonators with  $b = 695.5 \mu\text{m}$ ,  $Z = 25 \Omega$  and is presented in Figure 5.23. The second and third series correspond to the resonator with  $b = 193 \mu\text{m}$  and respectively  $Z = 30 \Omega$  and  $Z = 25 \Omega$  and are presented respectively at Figures 5.23 and 5.24.

Additionally, HB simulations are performed using the large signal model of the transistor, adapted to fit the de-embedded measurements, see Figure 5.8 to identify the oscillation frequency, phase noise and output power. The results of HB simulations are presented in Table 5.3. The phase noise values are on average close to  $-115 \text{ dBc/Hz}$  @ 1 MHz frequency offset from the carrier frequency for the LTCC cavities with  $b = 193 \mu\text{m}$  and to  $-118 \text{ dBc/Hz}$  @ 1 MHz frequency offset from the carrier frequency for the LTCC cavities with  $b = 695.5 \mu\text{m}$ .

Following Leeson's equation (5.1) the phase noise is analytically computed to be  $-117 \text{ dBc/Hz}$  @ 1 MHz frequency offset from the carrier frequency, using the characteristics of the LTCC resonators i.e. a loaded Q-factor on average  $Q_l = 27$ , results from Table 5.2. This evaluation assumes the active device noise figure at operating power,  $F = 4$  as known for the QUBIC4X technology, the oscillation frequency  $f_{osc} = 60 \text{ GHz}$ , the output power  $P_{out} = 7 \text{ dBm}$  and the offset frequency  $f_m = 1 \text{ MHz}$ .

Table 5.3.: Comparison between the small signal simulations, harmonic balance and measurement of the oscillators based on LTCC cavity resonators with different ratios  $o/w$  and heights.

| b<br>( $\mu m$ ) | $Z_c$<br>( $\Omega$ ) | $o/w$ | $f_{osc}$<br>(GHz) |         |       | Phase noise<br>(dBc/Hz@1MHz) |       |
|------------------|-----------------------|-------|--------------------|---------|-------|------------------------------|-------|
|                  |                       |       | Sim. ss            | Sim. HB | Meas. | Sim.                         | Meas. |
| 193              | 30                    | 1.5   | 59.92              | 59.94   | -     | -115                         | -     |
| 193              | 30                    | 1.6   | 59.90              | 59.83   | 59.91 | -115                         | -116  |
| 193              | 30                    | 1.7   | -                  | 59.73   | 59.97 | -114                         | -116  |
| 193              | 25                    | 1.1   | 60.05              | 60.02   | -     | -116                         | -     |
| 193              | 25                    | 1.15  | 59.98              | 59.93   | -     | -116                         | -     |
| 193              | 25                    | 1.2   | 59.93              | 59.84   | 59.98 | -115                         | -116  |
| 695.5            | 25                    | 1.3   | 60.25              | 60.24   | -     | -118                         | -     |
| 695.5            | 25                    | 1.4   | 60.24              | 60.19   | -     | -118                         | -     |

### 5.4.3 FC Assembly

The active MMIC chip is flip-chip (FC) mounted on top of the LTCC substrate, which integrates the cavity resonators, as presented in Figure 5.14. The FC is realized similarly to the MCM-D assembly using a FC150 flip-chip tool from Suss Microtec. The MMICs have been provided with gold stud bumps which are bonded to gold pads on the LTCC substrate by means of Au-Au thermo-compression bonding. The parameters of the FC bonding are as follows: at  $300^\circ C$  at the arm with MMIC and  $100^\circ C$  at the chuck supporting the LTCC substrate. The temperature and force is gradually increased during 3 min and maintained during 2 min. The applied bonding force per stud bumps is 300 gF per bump and corresponds to a total force applied of 5.2 KgF for 26 stud bumps. This Force is somehow relatively high as several MMICs were found to be cleaved after bonding. Additionally failures may have occur during the final assembly as the oscillators yield is quite low (3/20) as can be seen in Table 5.3. Unfortunately, no oscillator with a cavity resonator of a height  $b = 695.5 \mu m$  were functional. The top side of the MMIC, removed after FC assembly is presented in Figure 5.26. We believe RF shorting is happening between the stub bumps ( $50 \mu m$  diameter) placed relatively close to each other (pitch is  $150 \mu m$ ). Additionally, the stud bumps are enlarged due to the pressure of the flip chip and as it seems that metal is present in between two stud bumps.

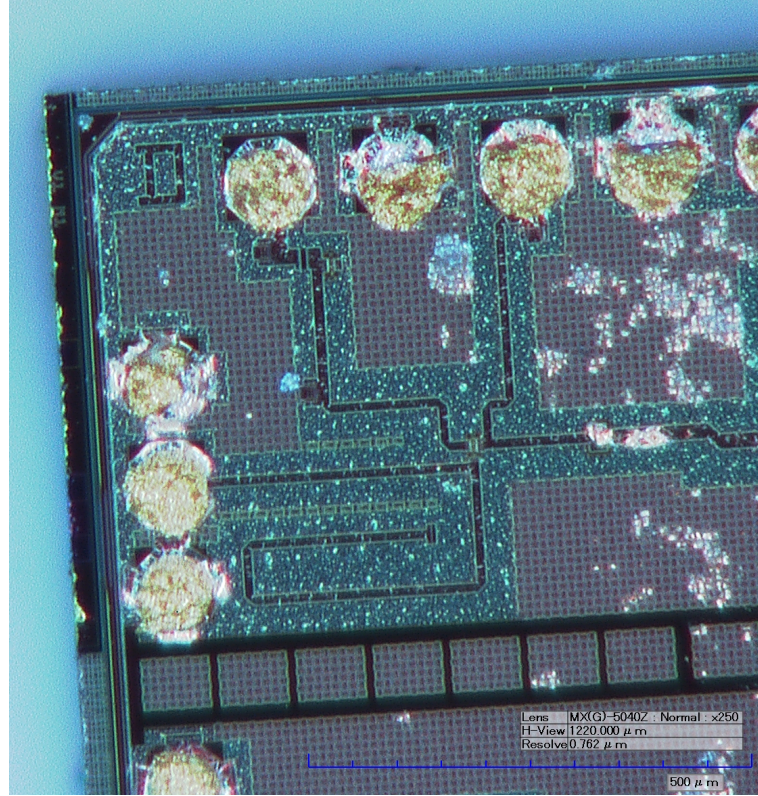


Fig. 5.26.: Top side of the MMIC, removed after FC assembly. RF shorting between the stub bumps could explain the low yield. Designed by Andreas John from AC Microwave GmbH, now with Hella KGaA Hueck & Co., realized by NXP Semiconductors in QUBIC4X technology [Deixler 04].

#### 5.4.4 Measurements

The measurements presented in Table 5.3, are addressing the frequency of oscillation, the output power and phase noise. Additionally, a sweep of the biasing condition: collector emitter voltage,  $V_{ce}$  is performed and these parameters are recorded. As an initial step, oscillations in a large frequencies band i.e. 40 to 80 GHz are checked. The oscillator's reflection coefficient is measured using an Agilent E8361A (PNA-X) Microwave Network Analyzer set in reception from 40 to 70 GHz. This band corresponds to the negative impedance delivered by the active circuit, presented in Figure 5.8. The only reflection coefficient above 0 dB is observed at 59.978 GHz (see Figure 5.27) and is identified with the oscillation frequency. For these measurements, the base emitter voltage,  $V_{be} = 1$  V, collector emitter voltage,  $V_{ce} = 2$  V and the drain current,  $I_d \approx 80$  mA. From Figure 5.27 we conclude that there are no other spurious oscillations at any other frequencies than the designed oscillation frequency, which is very close to 60 GHz.

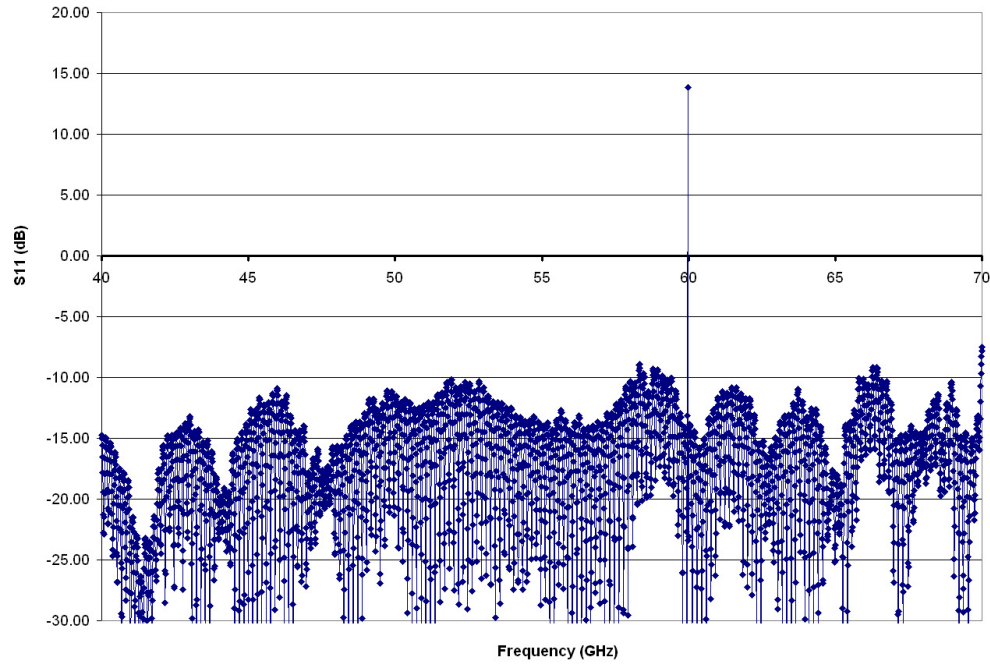


Fig. 5.27.: Reflection coefficient of the oscillator measured from 40 to 70 GHz with a PNA 67 GHz set in reception.

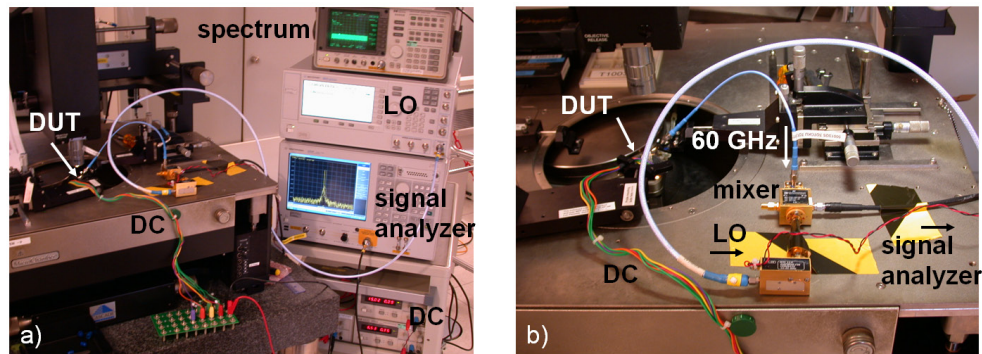


Fig. 5.28.: Phase Noise measurements set-up.

For the power measurements, the RF probes (1.85 mm pPROBE 67 GHz-GSG-150  $\mu\text{m}$  pitch) were connected via a 10 cm long RF coaxial cable, Totoku (M-M) 67 GHz to a coaxial to waveguide adapter Quinstar WR15 and finally to an Agilent V8486A V-Band power sensor along with a HP437B power meter. Following preliminary measurements, the loss of the coaxial cable is estimated to be 1.45 dB and the loss of the coaxial to waveguide adapter is estimated to be 1.3 dB. The 150  $\mu\text{m}$  pitch Ground-Signal-

Ground probes are characterized by an insertion loss of 1 dB at 60 GHz [Picoprobes]. Therefore a total loss of 3.75 dB is considered and added to the power meter indications: the measured maximum output power is +6.28 dBm, and the estimated output power is +10.03 dBm. The maximum power level obtained corresponds to a biasing:  $V_{be} = 1$  V,  $V_{ce} = 2.2$  V and  $I_d = 90$  mA which is around the designed bias condition:  $V_{be} = 1$  V,  $V_{ce} = 1.8$  V and  $I_d = 80$  mA.

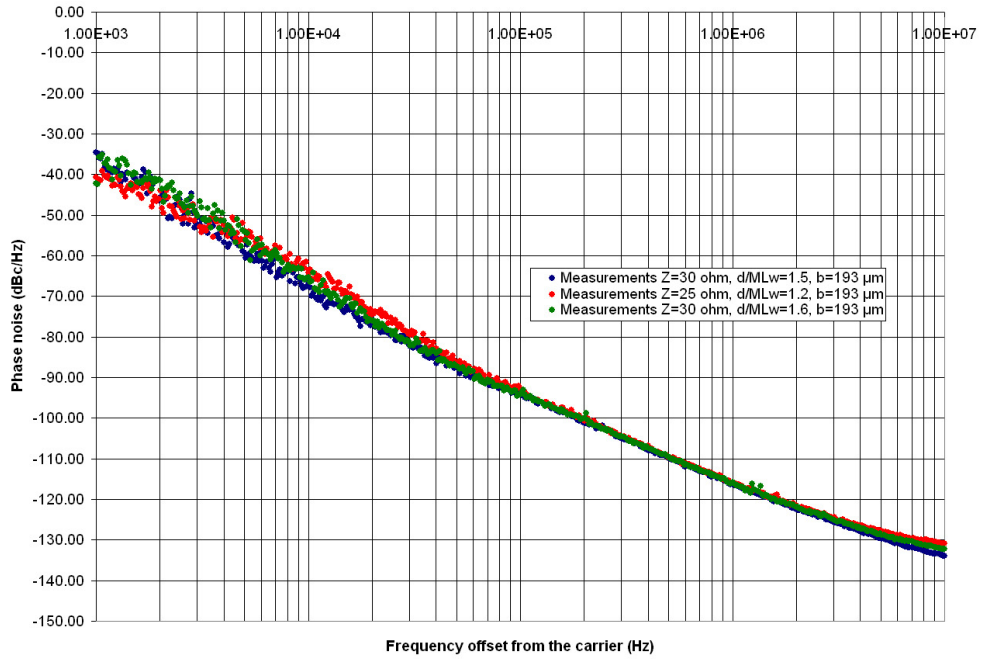


Fig. 5.29.: The Phase Noise of 3 different oscillators is consistently around -116 dBc/Hz @ 1 MHz offset from the carrier frequency.

For the Phase Noise measurement the signal analyzer Agilent 5052B was used. As this instrument is limited to 7 GHz, a down conversion scheme using an I/Q WR15 mixer from Ducommun with the RF generator Agilent E8257D 67 GHz, as local oscillator of which reference frequency was four times multiplied by a V-band active multiplier from Terabæam. The full measurement set-up is presented in Figure 5.28. The results are presented in Figure 5.29 for the three LTCC available oscillators presented in Table 5.3. A phase noise of -116 dBc/Hz @ 1 MHz offset from the carrier frequency is measured with consistency for the three oscillators.

Additionally the signal analyzer Agilent 5052B was used as spectrum analyzer, centered at the oscillation frequency with a 13 MHz span and 1001 measurement points. The spectrum results of the three LTCC oscillators presented in Table 5.3 are presented in Figure 5.30. The signal is analytically up converted, prior to the illustration in Figure 5.30 by adding the Intermediate Frequency, IF (variable around 240 MHz at the bias point  $V_{be} = 1$  V,  $V_{ce} = 1.8$  V) to the local oscillator frequency,  $f_{LO} = 14.932$  GHz. The oscillation frequency corresponds to the vector network analyzer measurement's,



presented in Figure 5.27 i.e.  $f_{osc} = 59.978$  GHz. As four times the local oscillator frequency is lower than the oscillation frequency, a reduction of IF corresponds to an increased measured frequency, as follows:  $IF = f_{osc} - 4 \times f_{LO}$ .

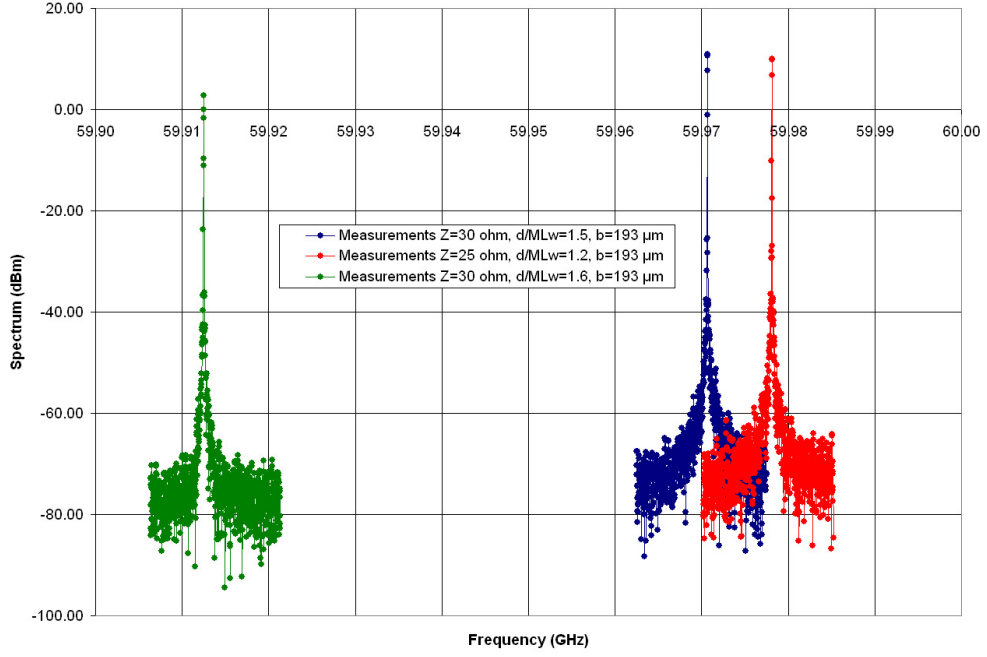


Fig. 5.30.: Spectrum of the 3 oscillators presented in Figure 5.29.

Typical I-V characteristics and time domain waveforms are presented in Figure 5.31, following parametric simulations, function of  $V_{ce}$ , realized with ADS. The  $I_c$ - $V_{ce}$  curves are superimposed with the steady state oscillation curves, for different operational points ( $V_{be} = 1$  V,  $V_{ce}$  swept from 1 to 5 V). The first curve of steady state oscillation corresponds to a common emitter configuration with a biasing:  $V_{be} = 1$  V,  $V_{ce} = 1.2$  V. The transistor is never operated in the saturation mode in any steady state oscillation. Until the collector-emitter voltage reaches  $V_{ce} = 2.8$  V, the transistor is operated in the cut-off mode and temporarily no current flows at the collector. This can be seen in the clipping of the  $I_c$ - $V_{ce}$  curves presented in Figure 5.31 and more synthetically in Figure 5.32 where the DC biasing currents are provided as a function of the biasing voltage  $V_{ce}$  with a maximum of the RF base current ( $I_{b_{RF}}$ ) is observed around 3 V.

In order to visualize the flow of the negative charges, the current is provided with a negative sign. The lowest simulated phase noise (HB in ADS) i.e. -119 dBc/Hz @ 1 MHz offset corresponds to this biasing ( $V_{be} = 1$  V,  $V_{ce} = 2.8$  V). The corresponding excursion of the RF I-V curve is maximal, as can be seen in Figure 5.31 and it is not clipped on the lower side by the cut-off mode. The oscillations corresponding to biases above  $V_{ce} = 2.8$  V, suffer from an upper side clipping due to the limitation of the out-

put collector current with increasing  $V_{ce}$ . The last steady state oscillation occurs for  $V_{ce} = 4.2$  V, close to the gain roll-off. The reverse bias of the collector-base makes the charge region of the junction to extend through the base all the way to the emitter-base junction. The transistor action is no longer possible, this is the punchthrough effect. The rapid increase in the collector current is similar to the avalanche breakdown [Sze 06].

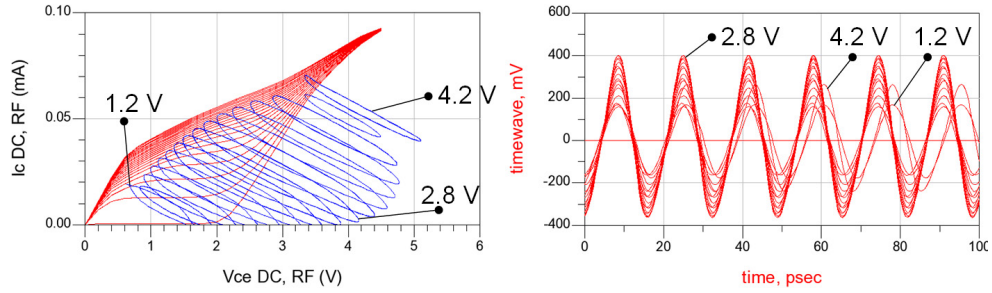


Fig. 5.31.: ADS simulations of the DC I-V characteristics and steady state oscillation curves as a function of the biasing voltage  $V_{ce}$ .

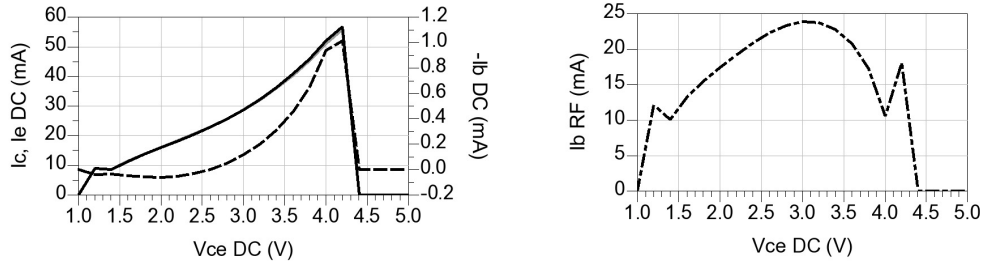


Fig. 5.32.: ADS simulations of the characteristic DC currents:  $I_c$ ,  $I_e$  and RF base current,  $I_{b, RF}$  as a function of the biasing voltage  $V_{ce}$ .

Similarly for the measurements, the output power is increased up to 10.03 dBm, as can be seen in Figure 5.34, for the biasing condition  $V_{be} = 1$  V,  $V_{ce} = 2.2$  V. For this biasing, the Phase Noise is -115 dBc/Hz @ 1 MHz offset from the carrier frequency  $f_{osc} = 59.99$  GHz, as presented in Figure 5.33 and the DC to RF efficiency,  $\eta_{DC/RF} = 5.09\%$ . Below  $V_{ce} = 1.9$  V, the output power is affected by the cut-off clipping of the  $I_c$ - $V_{ce}$  curve, as presented in Figure 5.31 for the simulations. Above  $V_{ce} = 2.3$  V, the output power is limited by the punchthrough effect. The optimal biasing condition for the Phase Noise is -116 dBc/Hz @ 1 MHz offset from the carrier frequency  $f_{osc} = 59.98$  GHz is  $V_{be} = 1$  V,  $V_{ce} = 2$  V, just in the middle of the power clippings, as can be seen from the sweeps in both  $V_{be}$  and  $V_{ce}$  presented in Figure 5.33. For this biasing the output power is 8.98 dBm, as can be seen in Figure 5.34 and the DC to RF efficiency,  $\eta_{DC/RF} = 4.94\%$ . Finally, the voltage sweep  $V_{ce} = 1.2$ -2.8 V

affects the oscillation frequency in a very limited way, as the maximal frequency shift is of 70 MHz, as can be seen in Figure 5.33. The frequency pushing is very low i.e. 44 MHz/V and is comparable to air micromachined cavity resonator oscillators [Song 99].

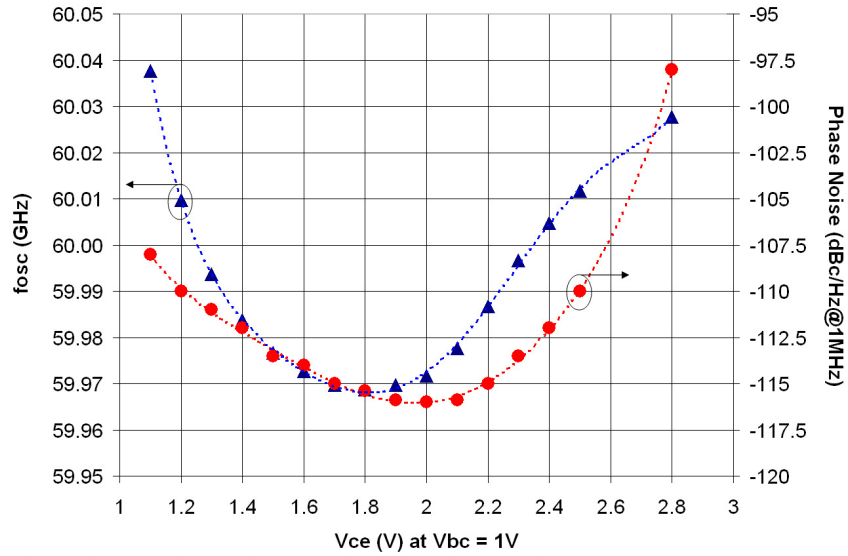


Fig. 5.33.: Oscillation frequency and Phase Noise in dBc/Hz @ 1MHz offset as a function of the biasing voltage  $V_{be}$ ,  $V_{ce}$ .

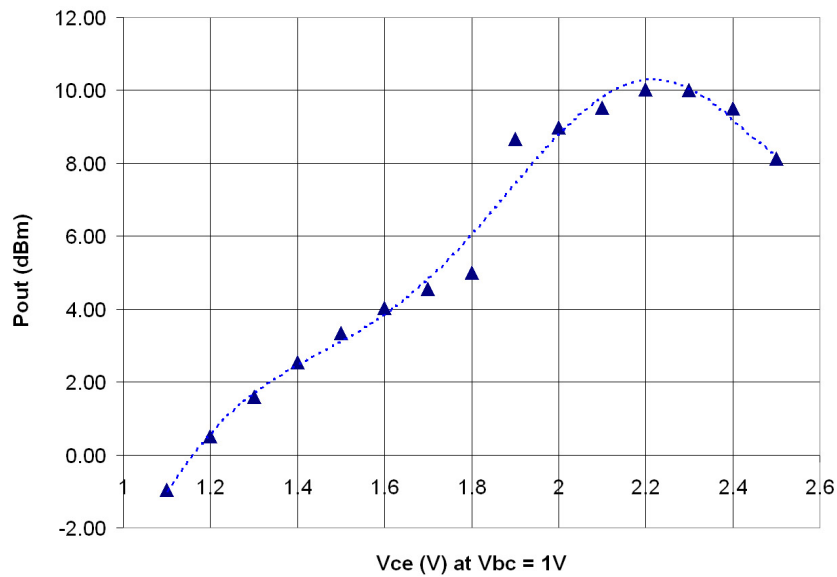


Fig. 5.34.:  $P_{OUT}$  of oscillator as a function of the biasing voltage  $V_{be}$ ,  $V_{ce}$ .



As can be seen in Figure 5.35, both active circuit and resonator are affected by a phase shift deviation from simulation of about respectively  $+15^\circ$  for the resonator and  $-17^\circ$  for the active circuit. Those figures come from reflection measurements performed before the FC assembly, on the resonator and active circuit at the negative resistance port. Those are reflection measurements and therefore a physical length of half the phase shift is practically responsible for the phase shift discrepancies i.e.  $+7.5^\circ$  and  $-8.5^\circ$ . As previously mentioned  $10^\circ$  corresponds to about  $100 \mu m$  of physical length considering the microstrip line used on MCM-D. The phase shift observed are caused by physical lines lengths around  $80 \mu m$ . Considering the stud bumps are having a total height of  $50 \mu m$ , any excessive compression during the FC assembly processing may result in an impact of the order of magnitude observed. The increased phase shift determined for the resonator may be due to a relative higher permittivity as resulting the presence of the MMIC die in close proximity and on top of the LTCC microstrip lines. Fortunately, the phase shifts of the active and resonator are in opposite directions and compensate each other which allows the oscillator to remain very close, in the best case up to 20 MHz close to the expected frequency of 60 GHz. Furthermore, as can be seen in Figure 5.33, the biasing conditions relative to  $V_{ce}$  are slightly tuning the oscillation frequency and 60 GHz can be obtained for  $V_{ce} = 1.2$  and 2.4 V. The latter biasing point is preferred as it allows a higher output power,  $P_{OUT} = 10$  dBm while preserving a fairly good phase noise performance of  $-113.5$  dBc/Hz @ 1 MHz offset from the carrier frequency. Finally, considering the same flip-chip assembly parameters, given the same active circuit, the resonator is required to provide a complex impedance ( $Z_R > 200 \angle 100^\circ$ ), to obtain a steady state oscillation at 60 GHz.

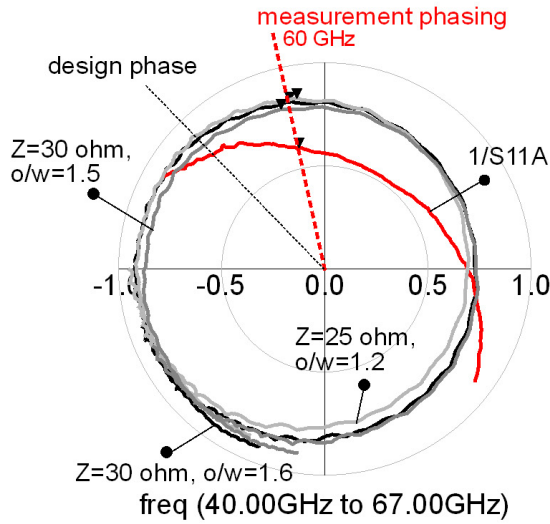


Fig. 5.35.: Combination of the reflection factors of the measured negative impedance amplifiers and resonators fulfilling the oscillation conditions around 60 GHz for the 3 oscillators presented in Table 5.3.

## 5.5 STATE OF THE ART OF INTEGRATED OSCILLATORS IN THE BAND 40-80 GHz

The evaluation of the Phase Noise, using Leeson's equation (5.1) of the oscillators realized in Sections 5.3-5.4 are presented in Figure 5.36 and placed in perspective of the expected phase noise as a function of their loaded Q-factor,  $Q_l$  and for different levels of output power i.e. 0 and 10 dBm.

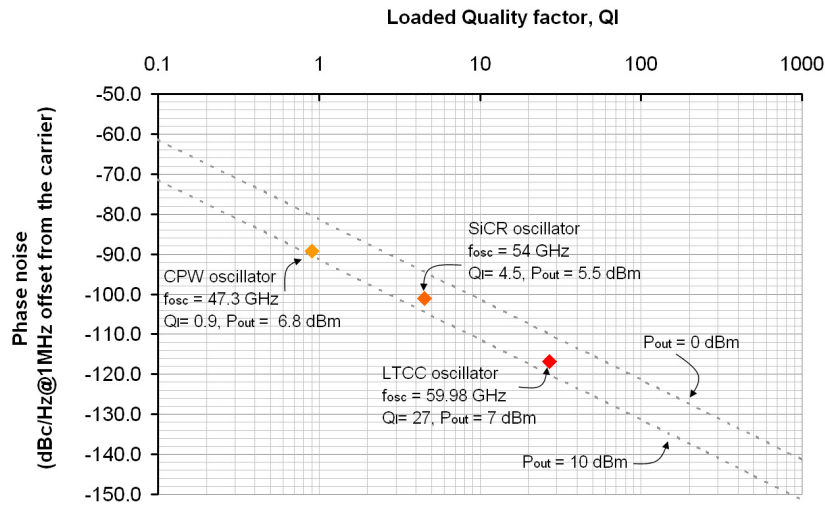


Fig. 5.36.: Characteristic oscillators realized within this thesis. Phase noise related to the loaded Q-factor, oscillation frequency and output power.

Accordingly to (5.1), higher the oscillation frequency, higher the phase noise and lower the performance *ceteris paribus*. The outlines in Figure 5.36 are provided for the active device's standard parameters: noise figure at operating power  $F = 4$ , oscillation frequency  $f_{osc} = 60$  GHz and offset frequency  $f_m = 1$  MHz. Qualitatively, it can be seen that at 1 MHz offset frequency from the carrier, the expected phase noise is -90 dBc/Hz for a  $Q_l = 1$ , -110 dBc/Hz for a  $Q_l = 10$  and -130 dBc/Hz, for a  $Q_l = 100$ . Around  $Q_l = 30$  the expected phase noise is about -120 dBc/Hz. As a reminder, the 3D $\mu$ Tune targeted a free-running 60 GHz Silicon micromachined cavity based oscillator with a phase noise of -136 dBc/Hz @ 1MHz offset from the center frequency with approximately 5% tuning range. Those specifications were perceived as very optimistic by the project reviewers and couldn't be achieved within the framework of the project, as a loaded Q-factor of  $Q_l = 300$  is required to provide a phase noise of -140 dBc/Hz, considering an output power of 10 dBm.

Table 5.4 summarizes the performance of the oscillators realized in this thesis, based on an common emitter SiGe QUBIC4X HBT. The comparison is based on the standard Factor of Merit (FoM) for oscillators, given in (5.3).

$$\text{FoM} = \mathcal{L}(f_m) - 20 \log \left( \frac{f_{osc}}{f_m} \right) + 10 \log \left( \frac{P_{DC}}{P_{OUT}} \right) \quad (5.3)$$

where  $f_{osc}$  is the oscillation frequency,  $f_m$  the offset frequency i.e. 1 MHz,  $P_{DC}$  the dissipated DC power, and  $P_{OUT}$  the output power.

Table 5.4.: Representative oscillators realized within the framework of this thesis, SiGe technology.

| Ref.                            | $f_{osc}$<br>(GHz) | Phase Noise<br>(dBc/Hz@1MHz) | $P_{OUT}$<br>(dBm) | $P_{DC}$<br>(mW) | FoM  | Topology                     |
|---------------------------------|--------------------|------------------------------|--------------------|------------------|------|------------------------------|
| CPW 820 $\mu m$                 | 47.3               | -95                          | 6.8                | 80.3             | -176 | CE<br>series feedback        |
| SiCRO<br>ML 580 $\mu m$         | 51.25              | -101                         | 3.5                | 85.0             | -179 | hybrid-CE<br>series feedback |
| SiCRO<br>ML 450 $\mu m$         | 54                 | -102                         | 5.5                | 90.1             | -183 | hybrid-CE<br>series feedback |
| LTCC Z=30 $\Omega$<br>d/w=1.6   | 59.92              | -115                         | 10.03              | 198              | -198 | hybrid-CE<br>series feedback |
| LTCC Z=25 $\Omega$ ,<br>d/w=1.2 | 59.98              | -116                         | 8.98               | 160              | -198 | hybrid-CE<br>series feedback |
| LTCC Z=30 $\Omega$ ,<br>d/w=1.5 | 59.97              | -116                         | 8.66               | 133              | -199 | hybrid-CE<br>series feedback |

The State of the Art in the band 40-80 GHz is presented per technology: heterojunction bipolar transistors (HBT) SiGe and III-V, in Table 5.5, high electron mobility transistors (HEMT) or heterostructure FET (HFET) and metal semiconductor field effect transistor (MESFET), in Table 5.6 and complementary metal oxide semiconductor (CMOS) RF circuits, in Table 5.7.

The topologies used with heterojunction bipolar transistors (HBT) in SiGe and III-V, presented in Table 5.5 are very diverse and we encountered, reflection types with both common emitter (CE) and base (CB), push-pull, differential and Colpitts oscillators. The lowest phase noise registered is provided by an InP single heterojunction bipolar transistor in a differential Colpitts topology, operating at 79.3 GHz and achieving -118 dBc/Hz @ 1 MHz offset from the center frequency [Lao 04]. At 60 GHz, until recently, the lowest phase noise of a free-running BiCMOS oscillator operating at 60 GHz achieved -104 dBc/Hz @ 1 MHz offset from the center frequency, in a differential Colpitts topology [Lee 04]. Improved results were provided by a push-push architecture using a dielectric resonator (DR) with a resonant frequencies at 29 GHz and achieving -112 dBc/Hz @ 1 MHz offset from the center frequency of 57.5 GHz [Sinnesbichler 00]. To date and to our best knowledge, the best phase noise performance at 60 GHz is -116 dBc/Hz @ 1 MHz offset from the carrier frequency  $f_{osc} = 59.98$  GHz is realized within the framework of this thesis. This is an hybrid architecture using LTCC cavity resonators in a common emitter series feedback topology. The characteristics of these

oscillators are presented in Figure 5.37 for the phase noise, in Figure 5.38 for the FoM (5.3) and in Figure 5.39 for the output power.

Table 5.5.: State of the Art of SiGe based oscillators in the band 40 to 80 GHz.

| Ref.               | $f_{osc}$<br>(GHz) | Phase Noise<br>(dBc/Hz@1MHz) | $P_{OUT}$<br>(dBm) | $P_{DC}$<br>(mW) | FoM  | Topology                  |
|--------------------|--------------------|------------------------------|--------------------|------------------|------|---------------------------|
| [Aoki 96]          | 63.9               | -95                          | 4                  | 120              | -174 | CE-series feedback        |
| [Baeyens 03]       | 74.0               | -91                          | 0                  | 170              | -166 | push-push Colpitts VCO    |
| [Baeyens 00]       | 84.0               | -80                          | -6                 | 78               | -154 | balanced Colpitts         |
| [Kim 03]           | 60.0               | -93                          | -4                 | 158              | -163 | CB-push-push VCO          |
| [Lee 05b]          | 60.0               | -77.5                        | -14                | -                | -    | CB-push-push              |
| [Lao 04]           | 79.3               | -118                         | -8                 | 95               | -188 | differential Colpitts     |
| [Li 02]            | 47.0               | -108.5                       | 5.6                | 280              | -183 | differential VCO          |
| [Li 03]            | 48.7               | -101.2                       | -2.5               | 490              | -166 | differential VCO          |
| [Lenk 03]          | 77.0               | -92                          | -2.3               | 486              | -161 | push-push                 |
| [Sinnesbichler 00] | 57.5               | -105                         | -8                 | -                | -    | DRO-push-push             |
| [Sinnesbichler 00] | 57.5               | -112                         | -14                | -                | -    | DRO-push-push             |
| [Sinnesbichler 00] | 58.0               | -108                         | 1                  | 160              | -182 | push-push                 |
| [Schwarz 99]       | 67.0               | -90                          | -15                | 86               | -152 | differential LC tuned     |
| [Wang 95]          | 62.5               | -104                         | 4                  | 40               | -188 | CE-series feedback        |
| [Wanner 06]        | 73.5               | -105                         | 3.5                | 140              | -184 | push-push                 |
| [Welch 06]         | 64.0               | -100                         | 5                  | 128              | -180 | CE-series feedback        |
| [Winkler 03]       | 63.6               | -87                          | -17                | 74               | -147 | differential Colpitts     |
| [Winkler 03]       | 76.1               | -91                          | -7                 | 128              | -161 | differential Colpitts     |
| [Floyd 05]         | 67.0               | -102                         | -8                 | 24               | -177 | differential Colpitts VCO |
| [Lee 04]           | 59.0               | -104                         | 4                  | 244              | -180 | differential Colpitts VCO |

The pHEMT, FET based oscillators, presented in Table 5.6 are dominated by a reflection type topology, with common source (CS), with the exception of a common gate (CG) topology demonstrating very good performance at 96 GHz with a phase noise of -101 dBc/Hz @ 1 MHz offset from the carrier frequency [Kirby 06].

Table 5.6.: State of the Art of pHEMT, FET based oscillators in the band 40 to 80 GHz.

| Ref.           | $f_{osc}$<br>(GHz) | Phase Noise<br>(dBc/Hz@1MHz) | $P_{OUT}$<br>(dBm) | $P_{DC}$<br>(mW) | FoM  | Topology              |
|----------------|--------------------|------------------------------|--------------------|------------------|------|-----------------------|
| [Caruth 00]    | 72.2               | -85                          | 3                  | -                | -    | CS                    |
| [Chen 03a]     | 62.0               | -78.3                        | 9                  | -                | -    | CS-series feedback    |
| [Chen 03b]     | 97.0               | -88                          | 0.7                | -                | -    | CS-series feedback    |
| [Funabashi 94] | 55.1               | -108                         | 3.7                | -                | -    | DRO-series feedback   |
| [Kashiwa 98]   | 55.9               | -103                         | 11                 | -                | -    | CS-series feedback    |
| [Kawasaki 95]  | 59.7               | -87                          | 9.1                | -                | -    | CS-series feedback    |
| [Kirby 06]     | 96.0               | -101                         | 0.4                | 45               | -185 | CG-series feedback    |
| [Song 09]      | 59.88              | -112                         | 9.7                | 102              | -197 | CRO-parallel feedback |
| [Wenger 93]    | 81.0               | -90                          | 0                  | 21               | -175 | DRO-series feedback   |

Table 5.7.: State of the Art of CMOS based oscillators in the band 40 to 80 GHz.

| Ref.             | $f_{osc}$<br>(GHz) | Phase Noise<br>(dBc/Hz@1MHz) | $P_{OUT}$<br>(dBm) | $P_{DC}$<br>(mW) | FoM  | Topology              |
|------------------|--------------------|------------------------------|--------------------|------------------|------|-----------------------|
| [Cho 05]         | 52.6               | -97                          | -16                | 27               | -161 | push-push VCO         |
| [Ellinger 04]    | 57.0               | -90                          | -6.8               | 21               | -165 | push-push VCO         |
| [Ellinger 04]    | 60.0               | -94                          | -17                | 9.6              | -163 | differential          |
| [Liu 04]         | 63.0               | -85                          | -4                 | 119              | -156 | push-push             |
| [Shigematsu 05]  | 52.5               | -86                          | -8                 | 41               | -156 | CS-series feedback    |
| [Tiebout 02]     | 51.6               | -90                          | -30                | 2                | -151 | differential          |
| [Wang 01]        | 49.5               | -99                          | -11                | 13               | -171 | differential          |
| [van der Wel 04] | 43.0               | -90                          | -14                | 7                | -160 | differential          |
| [Fong 04]        | 40.7               | -89                          | -8                 | 11.3             | -163 | differential          |
| [Kim 05]         | 44.0               | -101                         | -6                 | 7.5              | -179 | differential          |
| [Changhua 06]    | 56.5               | -79                          | -10                | 18               | -151 | differential          |
| [Huang 07]       | 61.0               | -108                         | -23.5              | 9.8              | -170 | double push-push      |
| [Koh 09]         | 55.77              | -106                         | -3.7               | 76.4             | -178 | CRO-parallel feedback |

The CMOS based oscillators rely essentially on push-push and differential architectures, topologies that adapt easily to the particular and typical digital design style where complementary and symmetrical pairs of p-type and n-type metal oxide semiconductor MOSFETs are used. An exception is mentioned to be a reflection type common source series feedback reported in [Shigematsu 05], with a quite low phase noise of -86 dBc/Hz @ 1 MHz offset from the center frequency (52.5 GHz). Accordingly to the Figure 5.39, no oscillators above 65 GHz are available in this technology.

The highest performance in terms of phase noise is provided in general by HBT oscillators, followed by FET and finally by CMOS, as can be seen in Figure 5.37. In general CMOS oscillators exhibit poor phase noise performance, on average 10 to 20 dB/Hz lower than HBT and III-V based oscillators. The high  $1/f$  phase noise of the transistor was evoked [Song 09] as a source of deterioration. Additionally, the lower loaded Q-factor passives of the CMOS technology limitations could be also removed by using hybrid air cavity resonators. This is shown in a realization where an air two-port micromachined cavity is placed in a parallel feedback architecture, using  $0.13\ \mu\text{m}$  CMOS active circuit [Koh 09]. With this topology, a phase noise level of -106 dBc/Hz @ 1 MHz offset from the carrier frequency of 55.77 GHz was demonstrated and is reported in Table 5.7.

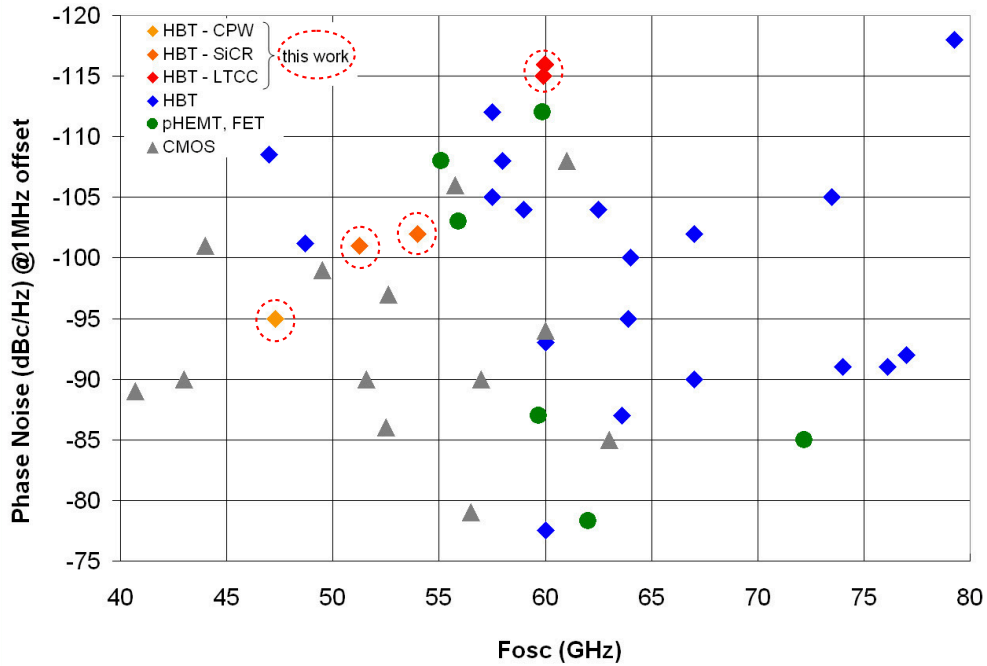


Fig. 5.37.: State of the art of integrated oscillators in the band 40-80 GHz. Comparison in regards to the Phase Noise versus the frequency of oscillation ( $f_{osc}$ ).

The results presented in Table 5.4 concerning the LTCC cavity resonator based oscillators are also the best in what concerns the FoM, as can be seen in Figure 5.38. The second best in terms of FoM is a GaAs pHEMT active stabilized by an air two-port micromachined cavity in a parallel feedback oscillator, with a measured phase noise of -112 dBc/Hz @ 1 MHz offset (59.88 GHz) [Song 09]. The extracted loaded Q-factor is mentioned to be  $Q_l = 352$  and for an output power of 6 dBm should have provided a phase noise of -139 dBc/Hz @ 1 MHz offset (60 GHz), accordingly to (5.1) and Figure 5.36. The lower phase noise performance is considered by the authors to be due to the excessive intrinsic  $1/f$  noise of the GaAs pseudomorphic high electron-mobility transistor and further improvements were expected by using a heterojunction bipolar transistors [Song 09]. Maybe a higher coupling in the oscillator operation, resulting in a lower loaded Q-factor could also contribute to the explanation of this lower performance.

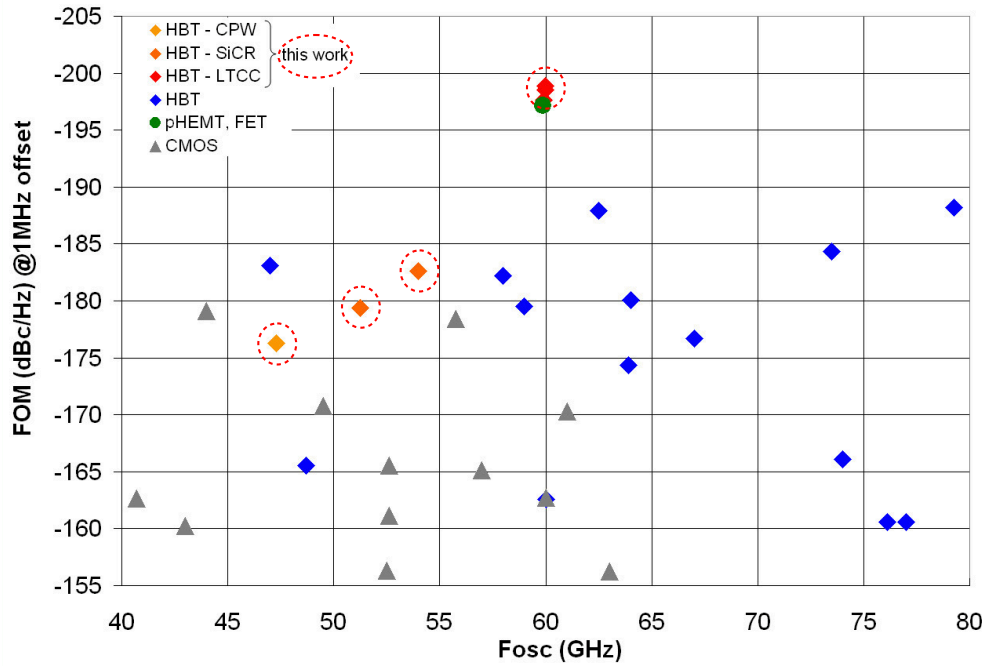


Fig. 5.38.: State of the art of integrated oscillators in the band 40-80 GHz. Comparison in regards to the Factor of Merit (FOM) versus. the frequency of oscillation ( $f_{osc}$ ).

The highest performance in terms of output power is provided in general by pHEMT devices but similar levels are obtained from SiGe. Additionally, common source or emitter topologies are providing the highest output power. Again, we must emphasize that the oscillator realized within the frame of thesis research is providing an output

power level amongst the highest ones known, up to date, i.e. around 10 dBm, as can be seen in Figure 5.39.

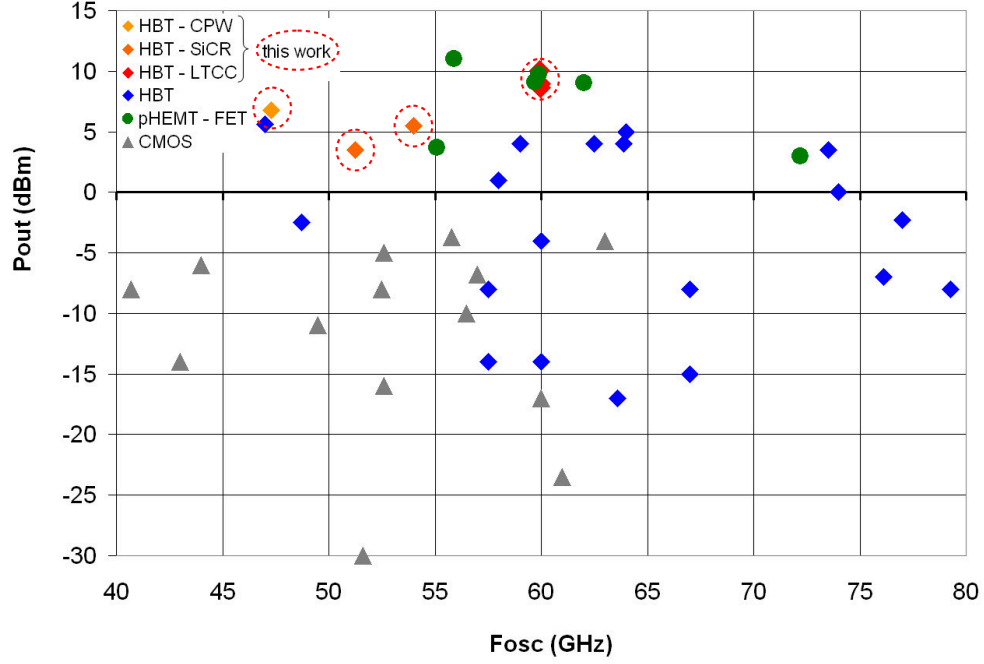


Fig. 5.39.: State of the art of integrated oscillators in the band 40-80 GHz. Comparison in regards to the output power ( $P_{out}$ ) versus the oscillation frequency ( $f_{osc}$ ).

## 5.6 CONCLUSIONS

The principle of operation of the oscillator is that an active circuit compensates the losses of the resonator joined in a virtual or physical closed loop circuit. The closed loop circuit amplifies noise in phase resulting in a magnitude oscillation.

Hybrid oscillators topologies, based on integrated waveguides offer much better performance as passives are characterized by low Q-factors at high operation frequencies. Those topologies consist in combining an active MMIC, generating a negative impedance to a higher Q-factor resonator realized in a different technology. At high frequencies, among the high performing resonators are the cavity resonators that may be filled or not with a dielectric. The filled resonators correspond to substrate integrated resonators in MCM-D and LTCC. The un-filled cavity resonators or air cavity resonators correspond to etched cavities soldered on top of a landing substrate. The latter are demonstrating higher performance in terms of Q-factors, as dielectric



losses are avoided while being larger in size (i.e. 2.4 reduction factor for LTCC; 3.4 for MCM-D). Additionally, these substrate integrated resonators do not require additional processing, nor assembly, provided the substrate vias are available.

There are different ways to design an oscillator, in our approach, we deliberately traded off the loaded Q-factor, by using higher coupling coefficients towards the fulfillment of the oscillation conditions, in terms of phase and magnitude:  $|\Gamma_A \Gamma_R| > 1$ . This approach is advised in order to avoid spurious oscillation at unwanted frequencies. The design of the oscillator is analyzed in terms of small-signal relations and S-parameters while harmonic balance simulations are required for the evaluation of the output power.

The best performing oscillator in the MCM-D technology demonstrated a phase noise of -102 dBc/Hz @ 1 MHz offset from the carrier frequency ( $f_{osc} = 54$  GHz). This oscillator in MCM-D demonstrated a 7 dB improvement when compared to an oscillator based on a CPW line resonator (i.e. a phase noise of -95 dBc/Hz @ 1 MHz) offset from the carrier frequency ( $f_{osc} = 47.3$  GHz). The improvement is due to the higher loaded Q-factor provided by the integrated cavity resonator in MCM-D over the CPW resonator,  $Q_{l\ SiCR}/Q_{l\ CPW} = 5$  and is in good agreement with Leeson's equation (5.1).

Unfortunately the MCM-D realization performed below performance in comparison, the LTCC implementation performed on target. This is not due to the technology itself but to the operation of the cavity resonator at the resonance frequency, which was failing in the MCM-D implementation. Considering similar RF characteristics for the integrated cavity resonators, the best performing oscillator in LTCC demonstrated a phase noise measured of -116 dBc/Hz @ 1 MHz offset from  $f_{osc} = 59.98$  GHz.

Furthermore, the thick LTCC cavity resonators presented, as expected better performance (i.e. loaded Q-factor,  $Q_l = 70$  for a LTCC cavity resonator with a height of  $695.5\ \mu m$ ) and therefore even lower phase noise figures were expected (i.e. -128 dBc/Hz @ 1 MHz offset from  $f_{osc} = 60$  GHz) but couldn't be demonstrated.

Within the design of the oscillator procedure it appears that the phasing of both active and resonator is of the highest importance. Additionally to the lumped model based simulations, de-embedded measurements are required to extract the right phasing at the connection port. Actually, all the parameters influencing the phase could not be determined in advance. Those parameters are the impact of the assembly technology and eventually the accuracy of the model at high frequencies, the technology was under development at the moment of design.

The optimal biasing for the phase noise performance is based on empirical characterization and corresponds to a biasing ( $V_{be} = 1$  V,  $V_{ce} = 2.2$  V) for the LTCC oscillator.

The results of the large signal analysis are helping to understand the non linear behavior of the oscillator. At lower biasing voltages, the oscillator is operated partially in cut off mode and temporarily no power is delivered to the oscillator, reducing the output gain. At very high collector biasing voltages, the punchthrough effect is increasing the depletion zone in the collector. As a consequence the base length is increasing and output current is reduced due to the increasing recombination in the longer base of the HBT.

The phase noise measurements were performed using a spectrum analyzer or an alternative method consisting in a down conversion scheme using a lower frequency band

signal analyzer. In this latest scheme, the output power was measured with a power meter. The simple measurements with the power meter are in any case advised as a control measurement.

Finally, it must be emphasized that the oscillators based on LTCC cavity resonators presented in this work, demonstrate the best performance as to our knowledge and to date, in what concerns both the phase noise and output power ( $P_{OUT} = 10.03$  dBm). This high performance contributed to the best FoM = -199 dBc/Hz @ 1 MHz offset, concerning the integrated oscillators in the band of interest (i.e. 40 to 80 GHz) while the oscillators oscillate exactly at the designed oscillation frequency i.e.  $f_{osc} = 59.98$  GHz. This achievement demonstrates the maturity of the LTCC technology, the robustness of the design technique and the potential of this hybrid topology.

Actually, increasing applications are demanding pure frequency synthesis (i.e. low phase noise) at very high frequencies. This topology is scalable (side dimensions of the cavity resonator), is completely integrated and presents very high performance at very high frequencies. Therefore this oscillator is an excellent candidate for high frequencies integrated oscillators. The high performance is due to the confinement of the fields within the cavity resonator, no substrate losses, nor excessive radiations nor spurious couplings were encountered.

## CHAPTER 6

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## CONCLUSIONS AND PERSPECTIVES

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### 6.1 INTRODUCTION

This research addressed the design, modelling and characterization of integrated cavity resonators for mm-wave. Additionally, the integration and miniaturization of such components and the tunability by internal perturbation was thoroughly investigated. Finally, applications based on these cavity resonator were developed. Those applications are substrate integrated filters and ultra-low phase noise oscillators.

### 6.2 CAVITY RESONATORS

Three parameters are fundamentally needed for the characterization of a cavity resonator: the unloaded Q-factor, the coupling coefficient and the unloaded resonance frequency, the other parameters can be derived. The characterization of cavity resonators is necessary, as numerous measurements realized were showing significant differences with the full-wave simulations, in terms of the Q-factor. This has a direct impact on the coupling coefficient and therefore on the impedance at resonance. Despite the discrepancy on the evaluation of the losses the evaluation of the resonance frequency is very accurately simulated. The differences between simulations and measurements observed are explained principally by the additional resistive losses at the melding of the cavity resonator, which is considered ideal in the HFSS simulations. Additional sources of loss are Ohmic, dielectric and radiative. Therefore I recommend

the normalization of the Q-factors, extracted from simulations by the ratio of the average Q-factors.

The coupling mechanism is also affecting the loaded resonance frequency, hiding the unloaded resonance frequency. This is a direct consequence of the energy stored in the feeding, which can be modelled by an additional shunt susceptance or in series reactance. As a consequence, the unloaded resonance frequency needs to be independently extracted. This is realized considering the detuning trend of a set of resonators, characterized by different coupling coefficients. Considering the extracted unloaded resonance frequency, lumped component models are realized. These model the RF response of the loaded cavity resonators as a function of physical parameters of the coupling structure. The detuning by the feeding could not be neglected and its impact on the complex impedance is required to be known quite precisely when resonators are used in oscillators. The coupling coefficient can be evaluated analytically considering the energy of transfer or considering the impedance, governed by the cavity's fields distribution. Practically, both transfer of energy and impedance are defining together the coupling coefficient.

Very high performance MCM-D integrated cylindrical cavity resonators at 60 and 75 GHz were demonstrated. Actually, typical unloaded Q-factors are respectively  $Q_0 = 117$  and  $Q_0 = 91$  at 60 GHz and 75 GHz. The cavity resonators are fully integrated in IMEC's MCM-D integration platform and no additional manufacturing is required, given the available interconnection TSVs. The overall dimensions of the resonators are respectively  $1.7 \times 1.7 \times 0.1 \text{ mm}^3$  and  $1.5 \times 1.5 \times 0.1 \text{ mm}^3$  at 60 GHz and 75 GHz, which represents an effectively low cost solution as the size on wafer is reduced due to the high permittivity substrate. It should be further emphasized that no substrate modes excitation were observed, which usually occur at such high frequencies. Those resonators are also suitable for oscillators, as the coupling coefficient can easily be made variable.

The realization of air cavity resonators was more challenging, as the standard capping is requiring a soldering temperature of  $300^\circ\text{C}$ , which is harmful for the cantilevers located under the capping. Adequate low temperature packaging technology was developed as the aluminum ( $\text{AlCu}_{0.5\%}$ ) MEMS cantilevers showed significant plastic deformation at  $100^\circ\text{C}$ . Those parameters improve both solder transfer and bond strength. The shear tests performed on the assemblies were resulting in an average shear pressures of 43 MPa, peaking at 78 MPa for the 2 kgF force and 7.5 s soldering time parameters at a temperature of  $240^\circ\text{C}$ .

A serious failure mechanism affecting the Q-factor of these resonators was identified. The investigation revealed that the Q-factor degradation is due a flow in the metallization of the cavity. Actually, this error occurred during the metallization by sputtering. The sputtered metal forms a gutter which shadows further a circular band of about  $2 \mu\text{m}$ . This results in an equivalent band discontinuity in the metallization. This discontinuity is responsible for the reduction of the Q-factor by a factor ten. Despite this, the tuning concept has been experimentally proven and separately very high Q-factor resonators have been demonstrated. The issues encountered were fully understood and turned into a valuable know-how towards the realization of such resonators. Therefore, further technological developments are required to achieve air cavity resonators with high Q-factor and reliable tuning actuators.

The LTCC technology shows similar RF functionalities as the MCM-D, but is different for the electric characteristics and for a technological aspect. This technology was also put at contribution for the elaboration of very high loaded Q-factor resonators, for oscillator design purposes and for large coupling coefficient cavity resonators for the design of a V-band Chebyshev seven-pole filter. This filter has a large-bandwidth (11.5%), measured low passband insertion loss (-2.4 dB) and high out of band insertion loss (-50 dB) in the lower stop band and (-30 dB) in the higher stop band. This is an integrated filter of compact size ( $2100 \times 9450 \mu\text{m}^2$ ) realized in a standard LTCC technology. The design of the filter is based on a filter synthesis method based on well established network methods and tabulated parameters. Those parameters are used to identify the electrical characteristics that are translated into physical dimensions, following an extraction of parameters from HFSS simulations. Additionally, a new feeding design was presented, meeting the large-bandwidth design requirements.

Finally, a new miniaturization concept for cavity resonators was presented, using high impedance surfaces. This concept opens the route towards the reducing in volume of cavity resonators, by a factor four. Acknowledging that a lot of satellites still embark heavy metallic based waveguides, this could serve cost reduction as the placement of satellites on orbit is prohibitive (\$10.000/kg). Nevertheless, the insertion of the HIS is reducing the overall Q-factor of the resonator and therefore a trade off between Q-factor performance and compactness should be considered i.e. in the idle case,  $Q_0 = 710$  at 7.9 GHz while it drops to  $Q_0 = 200$  at 4 GHz. However, measurements demonstrated a cavity resonator operating at 4 GHz with half the length and width of a standard metallic cavity resonator. This was successfully operated and proved the validity of the original concept.

### 6.3 CAVITY RESONATORS' TUNABILITY BY INTERNAL VOLUMES OF PERTURBATION

The investigation concerning the internal perturbation tunability of cavity resonators demonstrated that the impact of the perturbation shape located centrally (CPP) is lowering the resonance frequency while the dual shape (SAP) is increasing the resonance frequency. Additionally, the investigation of non-homogeneous volume of perturbation demonstrated that the variation of the height and of the lateral dimensions of the perturbation have both a controllable, evaluated and measured impact on the cavity resonator's resonance frequency. Therefore, the shape of the perturbation is optimized to perform the highest impact on the resonance frequency. On top of this, volumes of binary height (i.e. MEMS in up or down state) change the resonance frequency of the cavity resonator by an amplitude depending on their location in the cavity resonator. The small perturbation theory, based on the analysis of the field distribution inside the cavity resonator was used for this analytical study.

Integrated tunability of cavity-resonators is, to our knowledge, not demonstrated at 60 GHz elsewhere than in this thesis. The demonstration consists in a new tuning element for integrated air cavity resonators, namely the integration of a Faraday cage, shaped by upwards curved cantilevers, and defining a metallic enclosure inside the cavity resonator, tuning its resonance frequency. The upwards curved cantilevers ef-

fectively shield the air volume enclosed and therefore behave like a Faraday cage. The shielding of the array of cantilevers is equivalent to a volume with lateral dimensions equal to the array lateral dimensions and with a height equal to the tip deflection of the cantilevers. By controlling the cantilever tip deflection with a DC tuning voltage, the cage height is effectively varied resulting in a shift of the resonance frequency of the cavity resonator. This concept was successfully used to tune by 2% a 60 GHz silicon micromachined cavity resonator, provided the idle deflection of the incorporated MEMS i.e.  $30\text{ }\mu\text{m}$ . We measured a maximal frequency tuning of 1.2 GHz at 60 GHz with a tuning voltage of 16 V. The proof of concept of a tuning by the Side Annular Perturbation was also realized but the tuning range was much more lower. This is resulting from the partial actuation of several cantilevers among the array of cantilevers. The capping process is obviously affecting partially the capped cantilevers, by plastic deformation.

Nevertheless, the mm-wave packaging and the reliability of the MEMS involved in the actuation are antagonist. Even though a low temperature and fast capping process was developed, additional issues concerning the dielectric integrity during the actuation limits the reliability of this device. Actually, RF measurements regarding the expected Q-factors were unsatisfactory. The hypothesis to explain the degradation of the Q-factor by a disconnection of the metallization was verified and after correction of the process, a very high Q-factor resonator was finally realized at 74 GHz. Therefore, the investigation requires additional technological development to achieve in the same the realization of cavity resonators presenting high Q-factors, high tunability and high reliability. This represents a serious barrier to entry in the manufacturing of such components.

## 6.4 ULTRA-LOW PHASE-NOISE OSCILLATORS

An application driven investigation of an ultra-low phase-noise oscillators was realized. The ultra-low phase-noise oscillator realized is designed at 60 GHz and is based on a substrate integrated cavity resonator. This kind of realization is much welcomed as at mm-wave, as LC tanks are limited in performance and degrade oscillators' performance. Actually, the demand for such components is increasing as for example automotive RADARs are sought to be widely introduced by cars manufacturers. Similar devices, already on the market are available at prohibitive costs as their are implementing dielectric resonators, DR requiring manual placement and trimming.

The key issue in the design of ultra-low phase-noise oscillators is the phasing of both active and resonator parts. Therefore, additionally to the lumped models, measurements are required as all the parameters influencing the phase could not be determined in advance: the impact of the assembly technology, the accuracy of the model at high frequencies and the concomitant technological development. The design of the oscillators is easily analyzed in terms of S-parameters. In our approach, I deliberately traded off the loaded Q-factor, by using higher coupling coefficients towards the fulfilment of the oscillation's conditions in terms of phase and magnitude. This approach is advised in order to avoid spurious oscillations at unwanted frequencies. Leeson's equation is very good at predicting the oscillator's performance in terms of phase noise. The re-

sults of the large signal analysis are helping to understand the non linear behavior of the oscillator. At lower biasing voltages, the oscillator is operated partially in cut off mode and temporarily no power is delivered to the oscillator, reducing the output gain. At very high collector biasing voltages, the punchthrough effect is increasing the depletion zone in the collector. As a consequence the base length is increasing and the output current is reduced due to the increasing recombination in the longer base of the HBT.

The oscillators' performance is relatively good for the oscillator implemented in MCM-D. A phase noise of -102 dBc/Hz @ 1 MHz offset from the carrier frequency ( $f_{osc} = 54$  GHz) is demonstrated. Unfortunately this realization was not optimal and better performance was expected. Considering similar characteristics for the integrated cavity resonators in MCM-D and LTCC, the best performing oscillator in LTCC demonstrated a phase noise measurement of -116 dBc/Hz @ 1 MHz offset from  $f_{osc} = 59.98$  GHz. The improvement is not due to the technology itself but to the operation of the cavity resonator at the resonance frequency, which was failing in the MCM-D implementation. Furthermore, better performance was measured for the thick LTCC cavity resonators (i.e. loaded Q-factor,  $Q_l = 70$  for a LTCC cavity resonator with a height of  $695.5 \mu m$ ) and therefore even lower phase noise figure were expected as -128 dBc/Hz @ 1 MHz offset from  $f_{osc} = 60$  GHz, but couldn't be demonstrated because, most certainly of issues during the assemblage. The optimal biasing for the phase noise performance is based on empirical characterization and corresponds to a biasing ( $V_{be} = 1$  V,  $V_{ce} = 2.2$  V) for the LTCC oscillator.

Finally, it must be emphasized that the performance of the oscillator implemented in LTCC demonstrate the best performance as to our knowledge and to date, in what concerns both the phase noise and the output power ( $P_{OUT} = 10.03$  dBm) at 60 GHz. These are both contributing to the highest factor of merit, FoM = -199 dBc/Hz @ 1 MHz offset reported on integrated oscillators in the considered band of from 40 to 80 GHz. Additionally, the oscillator oscillates exactly at the designed oscillation frequency i.e.  $f_{osc} = 59.98$  GHz. This achievement demonstrates the maturity of the LTCC technology, the robustness of the design technique and the potential of this hybrid topology.

## 6.5 PERSPECTIVES

Very high performance was demonstrated by the integrated cylindrical cavity resonators at 60 and 75 GHz. Additionally, no substrate modes excitations were observed, which usually occurs at such high frequencies and for planar resonators. Therefore, these resonators represent very good candidates for the design of oscillators at very high frequencies, i.e. 94 GHz and above. Possible limitations could be encountered by the pitch of the vias, defining the resonators and to the degradation of the Q-factors, due to the dielectric losses. At such high frequencies, alternative modes of resonances could be used, i.e. the whispering mode, as this is offering higher Q-factors. Meanwhile, the small dimensions, inherent to such high frequencies are compensating the larger footprint, characteristic of this mode of resonance.

The air cavity resonators are characterized by very high Q-factors, nevertheless their footprint is quite large, even at 75 GHz. As these resonators are easily scalable towards higher frequency, the terahertz domain represents a promising niche. The lateral dimensions of a resonator, similar to the ones realized in this thesis, are  $a = d = 770 \mu\text{m}$  at 300 GHz;  $460 \mu\text{m}$  at 500 GHz and  $230 \mu\text{m}$  at 1 THz. Associated with Gunn, IMPATT and tunnel diodes they could form terahertz oscillators, potentially high power devices. If the dimensions become too critical, heterodyne architectures involving two different oscillators (push-push) could be used to up-convert the oscillation frequency while maintaining high output power levels. The latest is missing from the solutions available today.

Concerning the tunability by an integrated MEMS Faraday cage, the investigation of a dielectric free solution should be revisited. In such a solution, ground shunting bridges are used instead of the dielectric layer to isolate the DC biasing lines. This solution could alleviate technological issues related to the integration of the dielectric layer, between the cantilevers and the metallic bottom of the cavity. Such tunable resonators could further be used in oscillators and filters. For the multi-pole filters, a combinations of stacked cavity resonators could reduce the foot print of multiple pole filters and increase selectivity and stopband rejection. Additionally, during the characterization of the tunable cavity resonators, it was observed that the impedance is changing, affected by the fields' perturbation. As the oscillation conditions requires a well defined complex impedance, simultaneous impedance matching should be considered in future tuning investigations. Furthermore, as an extreme case, it was emphasized that already a thin post located in the center of a cavity resonator is having a considerable impact on the resonance frequency. Therefore, a single piezo-electric cantilever, provided with a large deflection and located in the center of the cavity resonator, could provide very large frequency shifts. Solutions implementing metals sustaining higher temperatures are also preferred for the realization of the MEMS.

The miniaturization concept could be broadened towards the investigation of different other kinds of symmetries, which in turn could increase further the level of miniaturization. The problem of the Q-factor reduction, observed during the experimentation should be further investigated in order to characterize the tradeoff between miniaturization and Q-factor reduction.



## APPENDIX A

### PRLC TO SRLC ABCD TRANSFORMATION

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In this appendix, we develop the transformation of the first Foster's model into the second one. The transformation of the loaded PRLC resonator in the SRLC resonator is using circuit theory, i.e. ABCD matrixes applied at each intermediated block.

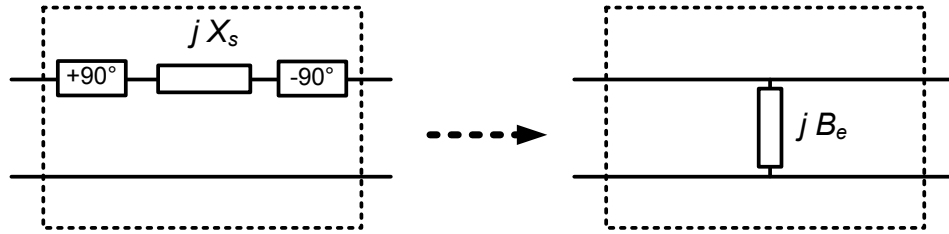


Fig. A.1.: Transformation of the reactance of the first Foster's form in to the susceptance of the second Foster's form.

$$Z = \begin{bmatrix} 0 & jR_c \\ \frac{j}{R_c} & 0 \end{bmatrix} \times \begin{bmatrix} 1 & jX_s \\ 0 & 1 \end{bmatrix} \times \begin{bmatrix} 0 & -jR_c \\ -\frac{j}{R_c} & 0 \end{bmatrix} \quad (\text{A.1})$$

$$Z = \begin{bmatrix} 0 & jR_c \\ \frac{j}{R_c} & 0 \end{bmatrix} \times \begin{bmatrix} \frac{X_s}{R_c} & -jR_c \\ -\frac{j}{R_c} & 0 \end{bmatrix} \quad (\text{A.2})$$

$$Z = \begin{bmatrix} 1 & 0 \\ \frac{jX_s}{R_c^2} & 1 \end{bmatrix} \quad (\text{A.3})$$

$$Y = j\frac{X_s}{R_c^2} \quad \text{and} \quad B_e = \frac{X_s}{R_c^2} \quad (\text{A.4})$$

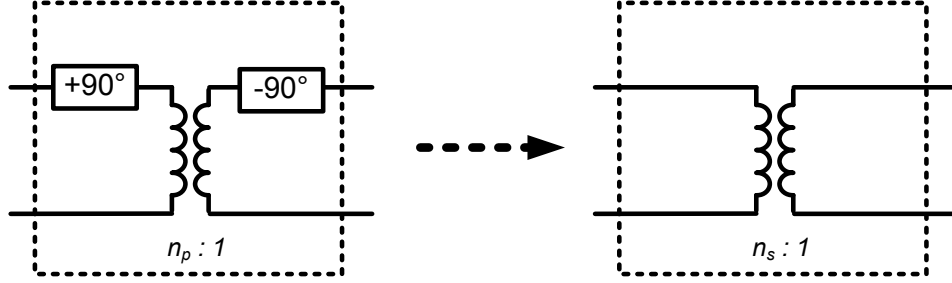


Fig. A.2.: Transformation of the transformer coefficient of the first Foster's form in to transformer coefficient of the second form.

$$Z = \begin{bmatrix} 0 & jR_c \\ \frac{j}{R_c} & 0 \end{bmatrix} \times \begin{bmatrix} n_p & 0 \\ 0 & \frac{1}{n_p} \end{bmatrix} \times \begin{bmatrix} 0 & -jR_c \\ -\frac{j}{R_c} & 0 \end{bmatrix} \quad (\text{A.5})$$

$$Z = \begin{bmatrix} 0 & j\frac{R_c}{n_p} \\ \frac{j}{R_c}n_p & 0 \end{bmatrix} \times \begin{bmatrix} 0 & -jR_c \\ -\frac{j}{R_c} & 0 \end{bmatrix} \quad (\text{A.6})$$

$$Z = \begin{bmatrix} \frac{1}{n_p} & 0 \\ 0 & n_p \end{bmatrix} \quad (\text{A.7})$$

$$n_s = \frac{1}{n_p} \quad (\text{A.8})$$

For the elements of the lumped resonator, the transformation is realized by identification considering the quarter wavelength transformation,  $Z = R_c^2 Y_p$ , where  $Y_p = \frac{1}{R_c}$  and  $R_c = 50 \, \Omega$ . This transformation is applied to the reactance of the parallel RLC circuit,  $Y_p = \frac{1}{R_p} + j\omega C_p + \frac{1}{j\omega L_p}$  and by identification with the impedance of the serial circuit  $Z_s = R_s + j\omega L_s + \frac{1}{j\omega C_s}$  is resulting in the following equations:

$$R_s = \frac{R_c^2}{R_p} \quad (\text{A.9})$$

$$L_s = R_c^2 C_p \quad (\text{A.10})$$

$$C_s = \frac{L_p}{R_c^2} \quad (\text{A.11})$$

## APPENDIX B

### CPP-SAP PERTURBATION ARRAYS

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The fabrication parameters of the Central Post Perturbation (CPP) and of the Side Annular Perturbation (SAP) cavity resonators are provided in Table B.1 and apply indifferently for CPP and SAP considering  $h_1 = b - H_{box}$  and  $h_2 = H_{box}$ .

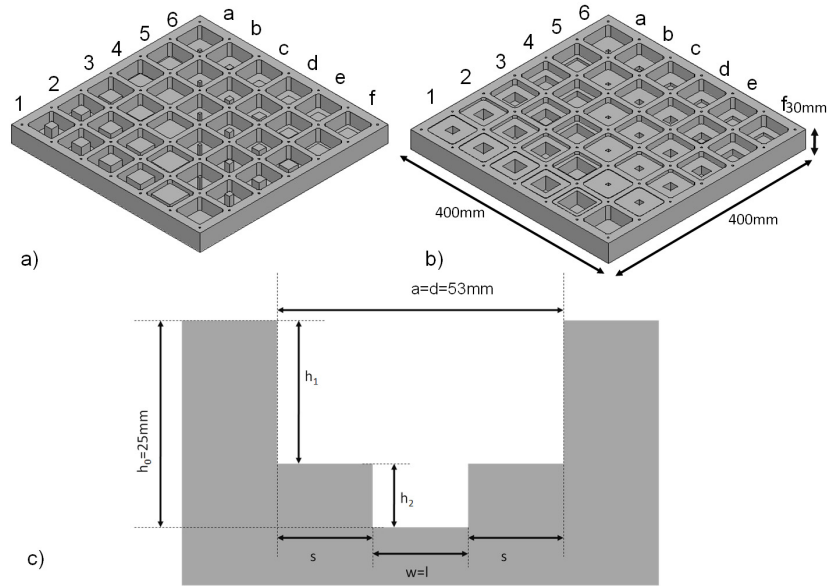


Fig. B.1.: Perspective view of the array of cavity resonators with a) Central Post Perturbation (CPP) and b) Side Annular Perturbation (SAP). c) Fabrication parameters definition referring to Table B.1.

Table B.1.: Fabrication parameters of the cavity resonators with CPP and SAP.

|    |          | Central Post Perturbation (CPP) |           |               |           | Side Annular Perturbation (SAP) |           |               |           |
|----|----------|---------------------------------|-----------|---------------|-----------|---------------------------------|-----------|---------------|-----------|
|    | location | $h_1$<br>(mm)                   | s<br>(mm) | $h_2$<br>(mm) | w<br>(mm) | $h_1$<br>(mm)                   | s<br>(mm) | $h_2$<br>(mm) | w<br>(mm) |
| 1  | a1       | 25                              | 24        | 5             | 5         | 20                              | 24        | 5             | 5         |
| 2  | a2       | 25                              | 21.5      | 5             | 10        | 20                              | 21.5      | 5             | 10        |
| 3  | a3       | 25                              | 19        | 5             | 15        | 20                              | 19        | 5             | 15        |
| 4  | a4       | 25                              | 16.5      | 5             | 20        | 20                              | 16.5      | 5             | 20        |
| 5  | a5       | 25                              | 14        | 5             | 25        | 20                              | 14        | 5             | 25        |
| 6  | a6       | 25                              | 11.5      | 5             | 30        | 20                              | 11.5      | 5             | 30        |
| 7  | b1       | 25                              | 6.5       | 5             | 40        | 20                              | 6.5       | 5             | 40        |
| 8  | b2       | 25                              | 24        | 10            | 5         | 15                              | 24        | 10            | 5         |
| 9  | b3       | 25                              | 21.5      | 10            | 10        | 15                              | 21.5      | 10            | 10        |
| 10 | b4       | 25                              | 19        | 10            | 15        | 15                              | 19        | 10            | 15        |
| 11 | b5       | 25                              | 16.5      | 10            | 20        | 15                              | 16.5      | 10            | 20        |
| 12 | b6       | 25                              | 14        | 10            | 25        | 15                              | 14        | 10            | 25        |
| 13 | c1       | 25                              | 11.5      | 10            | 30        | 15                              | 11.5      | 10            | 30        |
| 14 | c2       | 25                              | 6.5       | 10            | 40        | 15                              | 6.5       | 10            | 40        |
| 15 | c3       | 25                              | 24        | 15            | 5         | 10                              | 24        | 15            | 5         |
| 16 | c4       | 25                              | 21.5      | 15            | 10        | 10                              | 21.5      | 15            | 10        |
| 17 | c5       | 25                              | 19        | 15            | 15        | 10                              | 19        | 15            | 15        |
| 18 | c6       | 25                              | 16.5      | 15            | 20        | 10                              | 16.5      | 15            | 20        |
| 19 | d1       | 25                              | 14        | 15            | 25        | 10                              | 14        | 15            | 25        |
| 20 | d2       | 25                              | 11.5      | 15            | 30        | 10                              | 11.5      | 15            | 30        |
| 21 | d3       | 25                              | 6.5       | 15            | 40        | 10                              | 6.5       | 15            | 40        |
| 22 | d4       | 25                              | 24        | 20            | 5         | 5                               | 24        | 20            | 5         |
| 23 | d5       | 25                              | 21.5      | 20            | 10        | 5                               | 21.5      | 20            | 10        |
| 24 | d6       | 25                              | 19        | 20            | 15        | 5                               | 19        | 20            | 15        |
| 25 | e1       | 25                              | 16.5      | 20            | 20        | 5                               | 16.5      | 20            | 20        |
| 26 | e2       | 25                              | 14        | 20            | 25        | 5                               | 14        | 20            | 25        |
| 27 | e3       | 25                              | 11.5      | 20            | 30        | 5                               | 11.5      | 20            | 30        |
| 28 | e4       | 25                              | 6.5       | 20            | 40        | 5                               | 6.5       | 20            | 40        |
| 29 | e5       | 25                              | 24        | 24            | 5         | 1                               | 24        | 24            | 5         |
| 30 | e6       | 25                              | 21.5      | 24            | 10        | 1                               | 21.5      | 24            | 10        |
| 31 | f1       | 25                              | 19        | 24            | 15        | 1                               | 19        | 24            | 15        |
| 32 | f2       | 25                              | 16.5      | 24            | 20        | 1                               | 16.5      | 24            | 20        |
| 33 | f3       | 25                              | 14        | 24            | 25        | 1                               | 14        | 24            | 25        |
| 34 | f4       | 25                              | 11.5      | 24            | 30        | 1                               | 11.5      | 24            | 30        |
| 35 | f5       | 25                              | 6.5       | 24            | 40        | 1                               | 6.5       | 24            | 40        |
| 36 | f6       | 25                              | 26.5      | 0             | 0         | 25                              | 26.5      | 0             | 0         |

## APPENDIX C

### DE-EMBEDDING

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<sup>1</sup>Precise measurements are essential to characterize RF devices and to verify the accuracy of numerical simulations. To increase the accuracy of the measurements, conventional calibration techniques or de-embedding techniques exist. Among them, the most known are the TRL (Thru-Reflect-Line) / LRL (Line-Reflect-Line) and SOLT (Short-Open-Load-Thru) techniques. More details on different de-embedding techniques as LRM (Line-Reflect-Match) the LRRM (Line-Reflect -Reflect-Match) could be found in [Huynen 05].

In Figure C.1 is given a typical two-ports measurements problem. The DUT is embedded in systematic two-ports errors, mathematically represented by A and B.

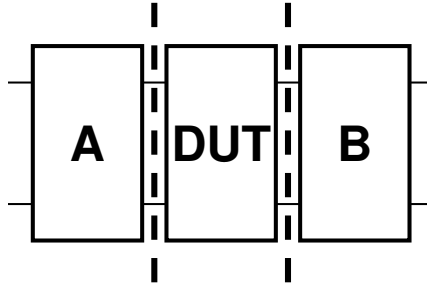


Fig. C.1.: Typical two-ports measurements problem. The DUT is embedded in systematic error, mathematically represented by A and B.

During the measurements, a vector network analyzer (VNA) is used to extract the four complex S-parameters of a two-port device under test (DUT). The VNA is connected to the DUT via RF coaxial cables and eventually probes or PCB connectors. The

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<sup>1</sup>This section is inspired by [Hayden 08].

length of these cables probes and connectors, as well as the effects of non-ideal termination can be removed by the calibration. Additionally, systematic errors, resulting from the reflection and transmission losses of the connectivity are removed.

The calibration consists in measuring different known calibration standards, measurements which are used in mathematical calculations to remove the measurement errors and to fix the reference plane. TRL is an advanced calibration, certainly the most popular. It is using a 8-term error model created by measuring the calibration standards: a thru (zero-length line), a line and a reflect that can be indifferently an open or a short. The thru may be replaced by a line (LRL). The error boxes A and B represent all the systematic errors and connectivity additional electrical lengths down to the center of the thru line, the initial measurement reference plane.

$$A = \frac{1}{S_{21}} \begin{bmatrix} -\Delta S & S_{11} \\ -S_{22} & 1 \end{bmatrix} \quad \Delta S = S_{11} \cdot S_{22} - S_{12} \cdot S_{21} \quad (\text{C.1})$$

Considering the thru matrix,  $T$  is the identity matrix, the measurement matrix of the thru standard  $M_T$  is given by the product of  $A$  and  $B$ , as follows:

$$M_T = A \cdot T \cdot B = A \cdot B \quad (\text{C.2})$$

The matrix,  $L$  accounts for the transmission line standard, which is inherently matched, since its impedance is the reference impedance of the errors. The line transmission behavior is set by the propagation constant and the physical length results in a diagonal matrix.

$$L = \begin{bmatrix} e^{+\gamma l} & 0 \\ 0 & e^{-\gamma l} \end{bmatrix} \quad (\text{C.3})$$

Considering  $S_{11} = S_{22} = 0$ , the propagation constant  $\gamma$  and the length of the line standard  $l$ , the reverse and forward transmission are reciprocal ( $S_{12} = S_{21} = e^{-\gamma l}$ ). The line measurement matrix,  $M_L$  is also given by the matrix product:

$$M_L = A \cdot L \cdot B \quad (\text{C.4})$$

Additionally, the matrix  $Q$  is computed as the product of the line and the inverse of the thru measurement.

$$Q \equiv M_L \cdot M_T^{-1} = A \cdot L \cdot B \cdot (A \cdot B)^{-1} \quad (\text{C.5})$$

$$= A \cdot L \cdot B \cdot B^{-1} \cdot A^{-1} = A \cdot L \cdot A^{-1} \quad (\text{C.6})$$

The matrix  $Q$  reduces to the inverse form  $ALA^{-1}$ , called a similarity transform of  $Q$ .  $L$  is a diagonal matrix with eigenvalues elements:  $p_{ii}$  of  $Q$ , as  $|Q - p_{ii} \cdot I| = 0$ .

The eigenvalues of  $Q$  are calculated by finding the roots of the determinant equation. The matrix  $A$  is a matrix made up of columns that are the eigenvectors  $v_i$  of  $Q$ , as  $|Q - p_{ii} \cdot I| \cdot v_i = 0$ .  $A$  is calculated using the roots of the matrix-vector product equation for each eigenvalue.

Finally, for the de-embedding of the DUT, matrixes  $A$  and  $B$  are required and following Figure C.1, the measurement of DUT is rewritten as follows:

$$M_{DUT} = A \cdot DUT \cdot B \quad (C.7)$$

The eigenvectors give the columns of  $A$ , but to multiplicative constants  $a_1$  and  $a_2$ .

$$A = [v_1 \quad v_2] \cdot \begin{bmatrix} 1 \\ a_2/a_1 \end{bmatrix} \cdot a_1 \quad (C.8)$$

Likewise,  $B^{-1}$  is known to the same multiplicative constants:

$$B^{-1} = M_T^{-1} \cdot A = M_T^{-1} [v_1 \quad v_2] \cdot \begin{bmatrix} 1 \\ a_2/a_1 \end{bmatrix} \cdot a_1 \quad (C.9)$$

Using this formalism, the  $a_1$  constant drops out leaving the complex unknown ( $a_2/a_1$  ratio) to fully determine the correction equation. This term is determined using the reflection measurement,  $M_R$ .

The corrected, de-embedded DUT is obtained from:

$$DUT = A^{-1} \cdot M_{DUT} \cdot B^{-1} \quad (C.10)$$

In this thesis, the TRL de-embedding technique was used to de-embed the measurements of the investigation of the 4 GHz cavity resonator. The feds were or microstrip or CPW line based. Therefore two sets of calibration standards were fabricated, simulated and measured.

The layouts of the standards for the microstrip line (Thru, Reflection, Line) are presented in Figure C.2-a), with also definition of the reference planes in -b) and in -c) picture of the fabricated standards. In Section 2.4.2, de-embedding was performed on the measurements at 4 GHz, concerning the investigation of cavity resonators' microstrip line feds. The results and details of the de-embedding are presented in Figure C.3. The details concerns the de-embedding which was performed on both HFSS simulations and measurements. Actually, for comparison a common reference plane was chosen i.e. the center of the circular aperture in the bottom PCB's metallization. The launching feature soldered to the microstrip line on PCB (coax to PCB) was not simulated in HFSS. Additionally, and for accuracy, the simulations required to take into account the feed line, which was de-embedded during the de-embedding process of the HFSS simulations realized using simulated standards.

The layout of the set of standards for the CPW line (Thru, Reflection, Line) are presented in Figure C.4-a), with also definition of the reference planes in -b) and in -c) picture of the fabricated standards. In Section 2.4.3 and 2.4.4, the de-embedding was performed on the measurements at 4 GHz, in the investigation of cavity resonators' CPW line feds with and without additional ending apertures. The results and details of the de-embedding are presented in Figure C.3 and Figure 2.4.4, respectively.

The de-embedding was performed on both HFSS simulations and measurements. Actually, for comparison a common reference plane was required to be chosen i.e. the short-ended CPW line, located at  $x_0/a = 0$  inside the cavity resonator. The launching features soldered to the microstrip line on PCB (coax to PCB) was not simulated in HFSS. Additionally, for accuracy, the simulations required to take into account the feed line, which was de-embedded afterwards.

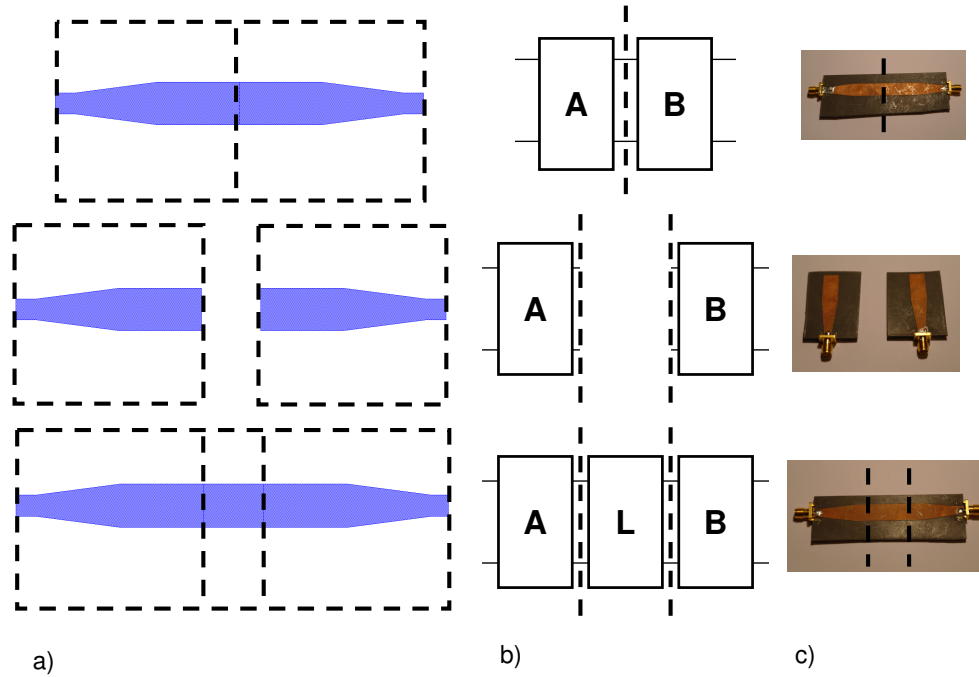


Fig. C.2.: a) Standards for the microstrip line (Thru, Reflection, Line) b) definition of the reference planes and in c) photo of the fabricated standards.



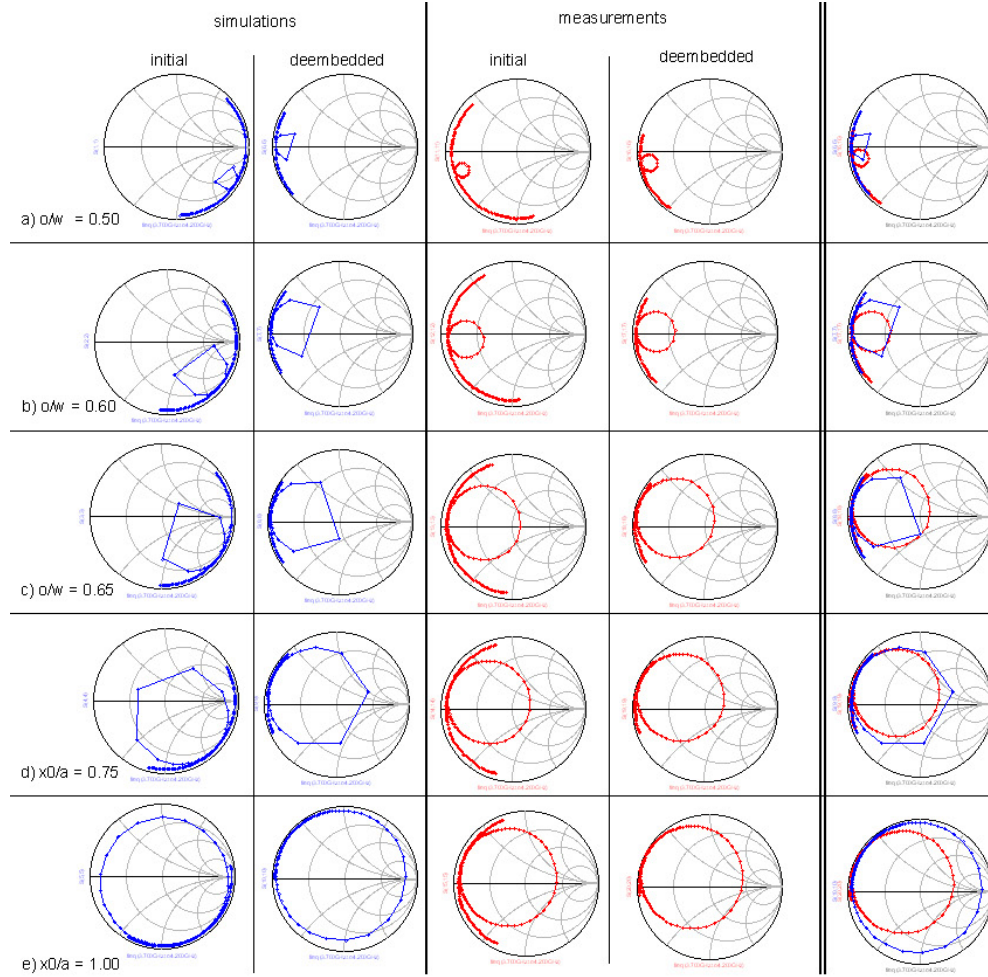


Fig. C.3.: The first column corresponds to raw HFSS simulations. The results of the de-embedding of the HFSS simulations are presented in the second column. The third column corresponds to raw measurement. The results of the de-embedding of the measurement are presented in the fourth column. Finally, the comparison between the de-embedded measurements and HFSS simulations of the ML fed cavity resonators are presented in the fifth column.

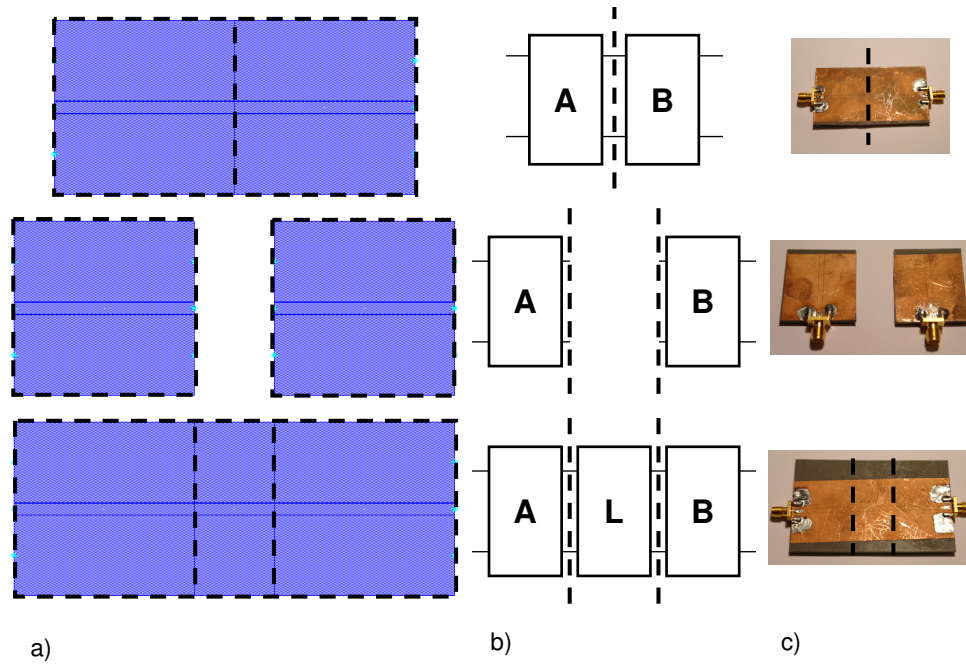


Fig. C.4.: a) Standards for the CPW line (Thru, Reflection, Line) b) definition of the reference planes and in c) photo of the fabricated standards.

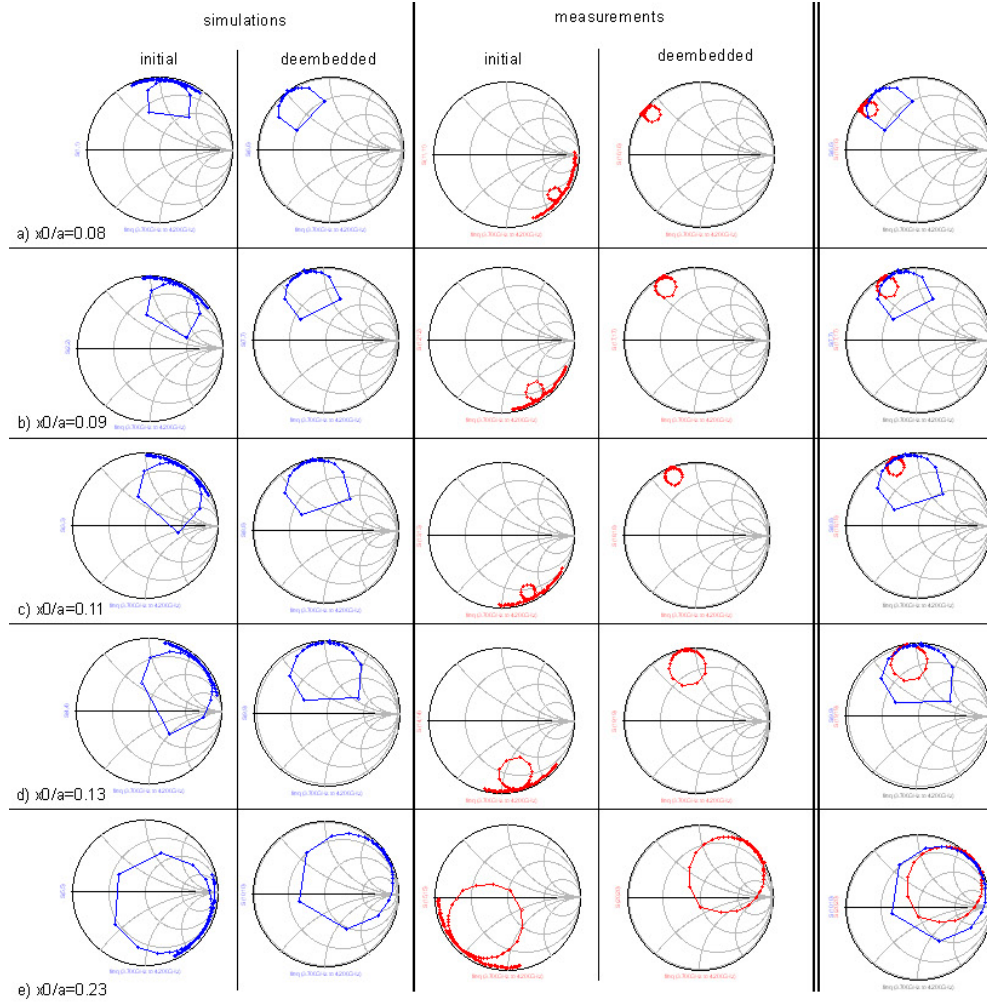


Fig. C.5.: The first column corresponds to raw HFSS simulations. The results of the de-embedding of the HFSS simulations are presented in the second column. The third column corresponds to raw measurement. The results of the de-embedding of the measurement are presented in the fourth column. Finally, the comparison between the de-embedded measurements and HFSS simulations of the CPW line fed cavity resonators are presented in the fifth column.

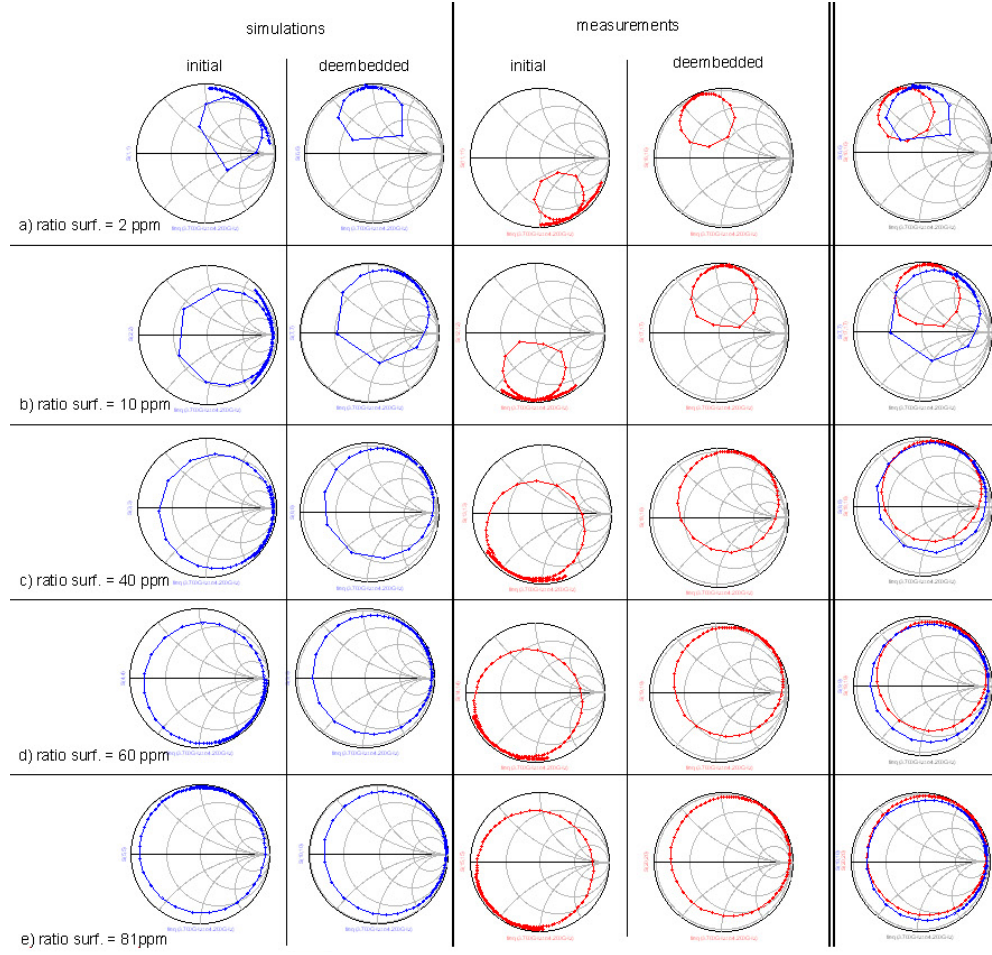


Fig. C.6.: The first column corresponds to raw HFSS simulations. The results of the de-embedding of the HFSS simulations are presented in the second column. The third column corresponds to raw measurement. The results of the de-embedding of the measurement are presented in the fourth column. Finally, the comparison between the de-embedded measurements and HFSS simulations of the CPW line fed, with additional apertures, cavity resonators are presented in the fifth column.

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## CURRICULUM VITAE

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Dragos Dancila was born in Bucarest, Romania in 1981. He graduated magna cum laude as an Electrical Engineer and was awarded a PhD in Applied Sciences from the Université catholique de Louvain, respectively in 2006 and 2011. He was granted a FRIA fellowship stipend by the FNRS, Belgium and pursued his PhD in strong collaboration with Imec. His research interests include mm-waves, ultra-low phase-noise oscillators, MEMS, metamaterials, RF components and systems. He received also a Complementary Master in Management from the Solvay Brussels School of Economics and Management in 2008.