Characterization and Reliability Studies Towards Piezoelectrically Actuated RF-MEMS Switches

Jiahui Wang

Propositions

Pertaining to the dissertation

Characterization and Reliability Studies Towards Piezoelectrically Actuated RF-MEMS Switches

by Jiahui Wang

1. RF-MEMS switches will become near-ideal once their reliability problems have been overcome (this thesis).
2. A simplified one-dimensional transducer model quantitatively expresses the coupling of electrical and mechanical behaviour of the RF-MEMS switches as long as the significance of the damping factor is weak (chapter 4).
3. Direct ion bombardments generate positively charged defects in PZT which are able to initiate dielectric breakdown under external electrical field stress (chapter 6).
4. Self-poling of PZT is at least partly caused by fabrication steps after the deposition of PZT (chapter 6).
5. The crystal structure and density of PZT largely determine the breakdown of metal-PZT-metal capacitors (chapter 7).
6. The ability of finding useful information is more important than learning new knowledge.
7. Independent judgment is a key characteristic of researchers - even ignoring negative reaction from others to this judgement is necessary.
8. Practicality is a criterion in choosing a research topic.
9. Researchers must make the impossible possible.
10. The goal can be achieved if people keep working towards it, even when the goal is poorly formulated.

These propositions are regarded as opposable and defensible, and have been approved as such by the promotor, prof. dr. Jurriaan Schmitz, and the daily supervisor, Dr. ir. Cora Salm.

CHARACTERIZATION AND RELIABILITY
STUDIES TOWARDS PIEZOELECTRICALLY
ACTUATED RF-MEMS SWITCHES

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CHARACTERIZATION AND RELIABILITY STUDIES TOWARDS PIEZOELECTRICALLY ACTUATED RF-MEMS SWITCHES

PROEFSCHRIFT

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Chapter 1

Introduction

The demands for switches have been greatly increased since the beginning of the Digital Age. Radio-frequency microelectromechanical systems (RF-MEMS) switches have become one of the most promising switches in communication equipment and satellite equipment, because of the developed fabrication technologies and the outstanding performance [1–5]. This thesis presents studies towards piezoelectrically actuated galvanic RF-MEMS switches. This chapter firstly explains why RF-MEMS switches become a feasible option in section 1.1, by introducing the development of MEMS technologies, research status and working principle of RF-MEMS switches. Then section 1.2 tells why RF-MEMS switches attract people’s attention by comparison with other RF switches. Performance criteria of RF-MEMS switches are expressed. The reliability problem is considered the main obstacle of the commercialization of RF-MEMS switches. Section 1.3 introduces the classification of RF-MEMS switches and explains why we aim to study piezoelectrical galvanic switches. Finally, the scope and outline of this thesis are presented in section 1.4 and section 1.5 respectively.

1.1 Introduction of RF-MEMS switches

The integration of electronics, optics, and mechanics on the micro-scale (or nano-scale) generates new approaches to applications, as illustrated in Fig. 1.1. Microelectromechanical systems (MEMS) devices integrate mechanical motion with electronics on the micro-scale [6]. Besides the information in Fig. 1.1, MEMS can also be integrated with magnetic elements such as MEMS magnetic actuators. MEMS technology can scale down to the nano-scale, also called nano-electro-mechanical systems (NEMS). The bulk micromachining techniques have been developed in the 1980s, using the

Figure 1.1: Illustration of the integration of electronics, optics, mechanics.

whole thickness of a Si wafer to build the micro-mechanical structures. The techniques enable high performance MEMS pressure sensors and MEMS accelerometers, which is a milestone in industry [4]. In order to combine MEMS and integrated circuits (ICs), surface micromachining techniques, which deposit sacrificial layers on the surface of a Si substrate, also became well developed after the 1980s and enabled MEMS in practice [4]. To be noticed, MEMS can use various substrate materials but the Si substrate is the most common one. The development of MEMS technologies makes it feasible to fabricate RF-MEMS devices (such as switches, resonators, waveguides, inductors and so on), which show growing importance for communications [6]. This thesis focuses on RF-MEMS switches which can achieve the near-ideal performance of mechanical switches and can be fabricated in micro/nano-technologies.

The main application areas of RF-MEMS switches include defense systems, intelligent base-station antennas, satellite switching networks, and high power phased arrays [7]. Many companies, such as Radant MEMS, Teledyne Scientific, HRL, IBM, NXP, EPCOS, Teravicta, Omron, Toshiba, and Advantest, have put extensive effort on research of RF-MEMS switches since 1990s [7]. However, only Radant [8] and Omron [9, 10] successfully have RF-MEMS switch products on the market. Recently, Advantest also claims to have RF-MEMS switch products being tested by customers [7]. Besides the companies, many universities and research labs also do research on the development of RF-MEMS switches. For example, Northeastern University fabricated the first prolonged galvanic RF-MEMS

Figure 1.2: Schematic cross-sectional view of RF-MEMS switches.

switches [7,11]; University of California - San Diego has been doing research on RF-MEMS switches for many years [1,12,13]; University of Limoges has been successful in wafer-level package on galvanic switches [14]. Thousands of papers have been published on RF-MEMS switches [7].

The RF-MEMS switches have two states: up-state and down-state, which are switched through the displacement of a movable electrode. Fig. 1.2 shows a schematic cross-sectional view of the RF-MEMS switches. In the up-state, the capacitance of the device is small which blocks the passing of the RF signal from the top to the bottom electrode; in the down-state, the device shows a large capacitance for capacitive switches and a small resistance for galvanic switches between two signal electrodes, which allows the RF signal to pass [1–5]. Various actuation methods can be used to displace the movable electrode of the RF-MEMS switches, as discussed in section 1.3.

1.2 Performance criteria for RF-MEMS switches

Compared with other RF switches, such as PIN diodes, pHEMTs and CMOS transistors, the MEMS switches have the following advantages.

Low insertion loss. The insertion loss is the power loss because of the insertion of an on-state (usually the down-state) RF-MEMS switch, which is equal to the difference between input and output power divided by in-

put power ($\frac{P_{in}-P_{out}}{P_{in}}$). The insertion loss of RF-MEMS switches can reach 0.1 dB up to 100 GHz [15].

High isolation. The isolation is the absolute value of the ratio of output power to the input power when a RF-MEMS switch is at off-state (usually the up-state). Because RF-MEMS switches are fabricated with air gaps, the up-state capacitance can be designed down to several femto-farad [15], providing excellent isolation higher than 20 dB [16].

Good linearity. The output voltage can be modeled as a Taylor series in terms of the input voltage [17]. For a device with good linearity, high order Taylor coefficients are negligible. The output power can linearly follow the low power input. However, this linearity does not hold when input power increases to a certain value (described by the 1 dB compression point) [3]. Nonlinear characteristics of devices can induce undesirable effects like gain compression or generation of spurious frequencies, resulting in increased loss, signal distortion, input interference among channels and so on [3, 17, 18]. Unlike semiconductor switches, the capacitance or resistance of RF-MEMS switches could be hardly influenced by RF signal, therefore, the RF-MEMS switches behave as extremely linear devices [1, 15, 19].

Low power consumption. Besides insertion loss, the switches also consume DC power. The power consumption describes the power dissipation at the working voltage of the switches, which is calculated by working voltage multiplied by current. For electrostatically actuated RF-MEMS switches, although the working voltage can reach 30 to 80 V, the current consumption is close to zero, thus the power consumption is also near zero [15].

Potential for low cost. The RF-MEMS switches can be fabricated by surface micromachining techniques; and can be integrated with various substrates, such as glass, quartz, polished ceramic, GaAs [1, 15]. Nowadays the MEMS switches are usually fabricated on Si substrates, because it can be integrated with the mature Si semiconductor technology thus allowing integrated circuits for digital, analog and RF mixed signal functionality to realize high level of integration [20].

Because of these advantages, especially the outstanding insertion loss and isolation, RF-MEMS switches may find wide application in switching networks for satellite systems, low-noise low-power circuits, portable wireless systems and so on [1]. For example, coaxial switches are widely used

in satellite applications because of their outstanding insertion loss and isolation, but they are heavy and expensive. RF-MEMS switches can easily meet the performance requirements and are much smaller and lighter than coaxial switches [1].

However, some disadvantages limit the application and even obstruct the commercialization of RF-MEMS switches, as described below.

Low switching speed. The switching time is the time that the RF-MEMS switches require to switch from one state to another state. It is determined by the dimensions, mass and squeeze film air damping, and typically amounts to microseconds or milliseconds [21]. Low switching speed is one of the main disadvantages of RF-MEMS switches compared with other semiconductor RF switches. Increasing actuation voltage helps to increase switching speed [21]. This is a trade-off carefully considered by designers because a high actuation voltage brings other problems.

High working voltage and dielectric charging problem. Typically, the RF-MEMS switches have a higher working voltage than the other RF switches. If the device contains a dielectric layer, a charging problem is likely to happen with the increase of duration of applied voltage stress on the dielectric layer, or with the increase of the number of repeated operations under high actuation voltage, as described in chapter 2 [12, 22–24]. Increasing the actuation area helps to reduce the actuation voltage, but this approach lowers the switching speed [21]. To be noticed, in case of high RF power application, the high actuation voltage is an advantage to prevent the RF power induced self-actuation, as explained in chapter 2.

Contact reliability. Repeated mechanical contact (in both capacitive switches and galvanic switches) and Joule heating (in galvanic switches) both degrade the contacts of RF-MEMS switches. Contact reliability is a main obstacle in the commercialization of RF-MEMS switches [1, 19, 20, 25–28]. Details of contact reliability are presented in chapter 2.

The qualitative comparison of performance criteria of various RF switches is shown in Table 1.1. The specific values strongly depend on the switch design, the actuation waveform shaping and the signal frequency, thus are not listed in the table. In short, the reliability problems are the main disadvantages of RF-MEMS switches [7]. Therefore, we aim to study on reliability aspects of RF-MEMS switches.

	MEMS switches	PIN diodes	FET
Insertion loss	Low	Medium	High
Isolation	High	Poor at low-end frequencies	Poor at high-end frequencies
Linearity	Perfect	Normal	Normal
DC power consumption	Low	Normal	Normal
Power handling	High	Low	Low
Switching speed	Slow	Fast	Fast
Working voltage	High	Low	Low
Reliability	Normal	Good	Good

Table 1.1: Comparison of performance criteria of various RF switches.

1.3 Classification

The RF-MEMS switches can be classified according to their actuation principle: electrostatic switches, electro-magnetic switches, piezoelectric switches and electro-thermal switches [12]. The classification can also be based on the actuation geometry, such as comb-drive actuators or gap-closing actuators. Nowadays, only the electrostatic and piezoelectric switches have shown reliable designs [7]. Electrostatic actuation has been widely studied. It uses the interaction force generated between two parallel metal plate electrodes to make the switch close. The electrostatically actuated RF-MEMS switches have relatively good reliability [7]. Electro-magnetic actuation could generate high force [29]. Compared with electrostatic actuation, piezoelectric actuation allows a lower operating voltage, a larger vertical open gap, microsecond operating speed, linear actuation, high energy density, and homogeneous stress distribution [19, 30, 31]. However, the processing and reliability of the piezoelectric material require more research, see e.g. [19, 32] and chapters 6 and 7 of this thesis.

The RF-MEMS switches can be divided into two groups by the device structure: galvanic (also called contact/ohmic) switches have a metal-metal (or conductor-conductor) contact, while capacitive switches have a metal-insulator-metal structure [16], as shown in Fig. 1.2. Compared with capacitive switches, galvanic switches are broadband and have a higher isolation [31]. Capacitive RF-MEMS switches are usually used above 6 GHz, whereas galvanic RF-MEMS switches are the only choice at 0.1 to 6 GHz and can also be used above 6 GHz [1]. However, the galvanic RF-MEMS

Figure 1.3: An envisioned piezoelectrical galvanic RF-MEMS switch.

switches have more reliability problems due to the contact of two metal electrodes, such as micro-welding caused by excessive contact heating, contact resistance increase caused by surface contamination and material transfer, etc. The reliability of galvanic RF-MEMS switches is also greatly influenced by the used RF power [20]. In addition, contact switches are sensitive to electrostatic discharges and electrical over-stress [20].

1.4 Scope

The EPAMO project, which supports the research described in this thesis, has the objective to explore the potential of unprecedented high density RF-MEMS switch arrays to be integrated in an energy-efficient agile RF transceiver with a reconfigurable antenna. This involves piezoelectric MEMS actuators based on Lead Zirconate Titanate ($\text{PbZr}_{1-x}\text{Ti}_x\text{O}_3$ or PZT) thin films. Fig. 1.3 shows an envisioned RF-MEMS switch of the EPAMO project [33]. The PZT layer serves as an actuator and the AlN layer serves as a passive structural material.

The main chapters of this thesis are divided into two main parts. The first part focuses on accurate characteristics of electrostatic capacitive RF-MEMS switches to get a more comprehensive understanding of RF-MEMS switches. A well-designed de-embedding/calibration and an accurate measurement is the basic for reliability study on RF-MEMS switches as well as their accurate modeling for circuit design.

As proposed in the project plan, we aim to change from electrostatic capacitive switches to piezoelectrical galvanic switches, because galvanic switches can be used in a wider frequency range than capacitive switches and piezoelectric switches have better performance in actuation voltage, speed, linearity etc. than electrostatic switches, as described in section 1.1. The reliability problem is the main disadvantage of RF-MEMS switches

in general, besides, piezoelectrically actuated galvanic switches suffer more serious reliability problems, especially the reliability of PZT thin film actuators, as described in section 1.2. That's why we aim to study PZT related reliability issues. Given the lack of industrial development samples of piezoelectrically actuated galvanic switches, the second part of this thesis focuses on the reliability of PZT thin films, which are potential piezoelectric actuators of our RF-MEMS switches.

1.5 Outline

This thesis starts with a literature review on various reliability issues on RF-MEMS switches in chapter 2. The main failure modes, namely stiction, contact resistance degradation, dielectric breakdown and dielectric charging, creep, and RF power induced self-actuation are expressed. Relevant reasons and solutions to these failure modes are discussed.

The first part of the main chapters of this thesis focuses on the accurate characterization of the electrostatically actuated capacitive RF-MEMS switches. Chapter 3 presents the origin of parasitics and provides calibration and de-embedding methods of the device under test (DUT). Five capacitance measurement methods are introduced and compared. The DUT capacitance is frequency independent across 11 orders of magnitude in frequency. However, the small-signal capacitance measured at low frequencies around pull-in and pull-out voltages significantly deviates from the classical capacitance-voltage (C - V) curves. Chapter 4 explains the deviations from the classical C - V curves by a one-dimensional transducer model at up-state. At down-state, the squeeze film effect influences the capacitance measurements, whereas it is absent in the experiments in vacuum. Chapter 5 discusses various approaches to determine the spring constant in the DUT. The results from the lowest vibration mode, from the pull-in voltage, and from the low-field C - V curve are compared, indicating that the pull-in voltage method might be the best suited for in-line process control.

The other part of the main chapters of this thesis is the reliability study of PZT thin films, which will be used as the piezoelectric actuators of the EPAMO RF-MEMS switches. The PZT properties are influenced by fabrication processing especially plasma etching. Chapter 6 compares two kinds of metal-PZT-metal capacitors, only one of which contains PZT directly bombarded by plasma particles. Both the experimental results as well as theory indicate that continuous direct ion bombardments decrease PZT permittivity and worsen PZT reliability. Chapter 7 aims to study the physics of PZT degradation and breakdown, by doing ramped voltage stress

(RVS) and time-dependent-dielectric breakdown (TDDB) measurements on unpackaged metal-insulator-metal (MIM) PZT capacitors with various stacks in various measurement conditions.

Finally, the thesis ends with the conclusions in chapter 8.

Chapter 2

Overview of RF-MEMS switches and PZT reliability

This thesis aims to have a more comprehensive understanding and work on the reliability of RF-MEMS switches, and study towards piezoelectrically actuated galvanic RF-MEMS switches (as described in chapter 1). This chapter reviews the publications of reliability on RF-MEMS switches. For galvanic switches, the dominant reliability problem relates to the metal-metal contact surfaces. Stiction and contact resistance degradation are the two main failure modes. For capacitive switches, the main reliability problem relates to the dielectric charging. To be noticed, stiction may also happen to capacitive switches; and the dielectric breakdown/charging also happen to galvanic switches when a dielectric layer involved (such as PZT film in galvanic switches). Since the reliability of piezoelectric material is the key point of succeeding in piezoelectric RF-MEMS switches, the reliability of PZT thin film actuators are reviewed in chapter 2, providing background knowledge for the studies in chapter 6 and 7.

This chapter starts with the main contact reliability issues: stiction in section 2.1 and contact resistance degradation in section 2.2. Then the dielectric charging/breakdown issues are reviewed in section 2.3. Section 2.4 goes towards piezoelectric switches, treating the reliability of PZT thin films. Last but not least, section 2.5 briefly introduces another two failure modes: creep and RF power induced heating/self-actuation.

2.1 Stiction

Stiction (or adhesion) is one of the most widely observed failure modes in MEMS devices. It means that the movable electrode of the RF-MEMS

Figure 2.1: Schematic contact surfaces with a certain roughness. [37]

switch does not restore to its rest ‘up’ position after removal of the actuation stimulus. Stiction is caused by an enhanced adhesion force between the two contact surfaces. The adhesion force is a tensile force needed to be overcome to separate the two contact surfaces, including capillary force, van der Waals forces and electrostatic force between the surfaces [34]. Stiction will happen, when the adhesion force is larger than the spring force even when the actuation voltage is removed. This section describes the possible reasons causing stiction and the corresponding solutions to mitigate stiction.

2.1.1 Contact shape evolution

The repeated mechanical contact and separation may induce plastic deformation of the contact surfaces, which is called contact shape evolution or contact wear [35, 36]. The contact surfaces cannot be absolutely smooth but have a certain roughness. Only part of the contacting surfaces come into real contact, as shown in Fig. 2.1. The contacting spots are known as ‘a’-spots and represent the real contact area. Contact shape evolution greatly influences the surface roughness and the real contact area, thus influences the adhesion force. Contact shape evolution also causes contact resistance change as discussed in section 2.2.

The magnitude of adhesion force strongly depends on the real contact area [34]. The two main contributions of adhesion force are capillary force and Van der Waals force. The capillary force is induced by attraction between water dipoles on contact surfaces. The Van der Waals force is induced by attraction between permanent dipoles or corresponding induced dipoles on contact surfaces. The force between dipoles decreases rapidly as the distance between the dipoles increases [37]. In other words, the surfaces outside the real contact area are far away from each other and have negligible contribution to the adhesion force. The adhesion force is

Figure 2.2: A typical stress-strain curve.

determined by the real contact area [37]. For the same surface materials and surface conditions, the adhesion force increases with the real contact area.

The applied load on the contact surfaces is called impact force. When the asperities on both contact surfaces touch, their deformation creates a force to balance the external impact force. Fig. 2.2 shows a typical relation between stress (force) and strain (deformation). The recoverable elastic deformation involves stretching of atomic bonds in the material, but the atoms do not move past each other [37]. In elastic region, the stress-strain curve is linear, the slope of which equals to the Young's modulus of the material. When the stress exceeds a threshold value (yield strength), the permanent plastic deformation happens. Some atomic bonds are broken because of the movement of dislocation inside the material lattice [37].

If the contact loading induces only elastic deformation, the unloading will also be elastic and the adhesion force will be constant over time (i. e. no degradation occurs). If the loading includes some plastic deformation, the electrode material can respond in two manners: the ductile mode and brittle mode, as shown in Fig. 2.3. The ductile separation mode occurs within the softer one of the contact surfaces when the average tensile force becomes equal to the hardness of the softer surface [35]. Material transfer occurs in the ductile mode. The material of the softer surface may attach to the harder surface. The asperities of the contact surfaces will change. The adhesion force will decrease if the asperities grow. The adhesion force will increase if the asperities shrink. The brittle separation mode occurs at the interface of the contact material and the brittle plastic deformation makes the surfaces smoother, therefore, the adhesion force increases [35]. The changes at contact surfaces mentioned above can be observed by a real-time observation technique in [25]. In conclusion, the adhesion force is

Figure 2.3: Relationship of the loading force, the contact shape evolution and the adhesion force. [35]

constant for elastic loading, however, it depends on the maximum loading force for plastic loading.

There are two ways to reduce the contact shape evolution. One is to reduce the impact force. Jain et al. point out that the high-velocity hard landing of the movable contact on the fixed contact is a main reason of the large impact force. The impact force can be greatly reduced in case of low-velocity soft landing, by using the fractal surfaces of the movable and fixed contacts [38]. McGruer et al. mention that using a specially tailored actuation pulse may significantly reduce the impact force [35]. Czaplowski et al. propose a soft-landing waveform, which consists of an actuation pulse (larger than pull-in voltage), a coast time at 0 V and a hold voltage (slightly larger than pull-out voltage) [39, 40]. The movable electrode gets kinetic energy to contact during the actuation pulse; the restoring force and damping slow the movable electrode to near-zero impact velocity during coast time; the hold voltage maintains the switch in contact-state.

The other way is to reduce the contact shape evolution is to use harder electrode materials. The threshold of force which induces a plastic deformation depends on the modulus of the contact materials. If a harder contact material is used, less plastic deformation and material transfer will happen during loading, and may result in smaller adhesion force during unloading. For example, McGruer et al. show that gold has smaller hardness and larger adhesion force than ruthenium [35].

Beside the above solutions, researchers also improve the mechanical design of the switch to increase the restoring force [22]. This method may help to mitigate stiction caused by various reasons. An increased restoring force can be achieved by optimizing the thickness, length or shape of the mov-

able cantilever or membrane [22, 41]. The trade-off between the actuation voltage and the restoring force should be carefully considered [22].

2.1.2 Capillary effect

Water molecules are usually present at any surfaces because of spontaneous fluid condensation [34, 36]. The formation of small liquid bridges between two hydrophilic surfaces is called capillary condensation. Capillary force is the inter-molecular attraction force because of capillary condensation [42]. The water molecules are dipoles. Opposite charges in the water dipoles can form hydrogen bonds. The force of hydrogen bonding is weaker than the covalent bonding force but is much stronger than the van der Waals force. Therefore, the capillary force originating from the dipole attraction between water molecules significantly increases the adhesion force [34].

A humid environment increases the concentration of water molecules and hence may increase the capillary condensation and the capillary force. Therefore, capillary effect induced stiction is easy to happen in the presence of moisture [36, 37]. In addition, the capillary effect induced stiction will be a more significant problem with the decrease of device size because the restoring force (the elastic force of the spring) usually decreases faster with device size than the capillary force [36, 37].

There are many ways to mitigate the influence of the capillary force. One solution is to use a hydrophobic coating on the contact surface. For example, a self assembled monolayer (SAM) coating could be used to tailor the surface energy and terminate specific bond, making the surfaces hydrophobic [34]. However, for a galvanic RF-MEMS switch, the contact surface must be covered with a metal with low resistivity, thus this solution is not always practical [37].

The second solution is to use the device in a water-free environment. Hermetic packaging may reduce humidity around the encapsulated devices. However, the packaging has to be carefully designed to prevent organic contaminations from encapsulation parts [1, 34]. To be noticed, wafer-level measurements are required in our studies (both for accurate characterization of RF-MEMS switches in chapter 3 to 5 and reliability of PZT thin films in chapter 6 and 7). Therefore, we cannot package the devices under test, instead, we heat the sample in dry air flow to get rid of the humidity, as described in chapter 3 and chapter 7.

The third solution is to remove the electronegative atoms from the surface. For example, a careful final etching process on the Si surface with ammonia fluoride (NH_4F) could remove the oxygen [34].

The fourth solution is to increase the surface roughness. As described in

2.1.1, the capillary force increases with the real contact area. Increasing the surface roughness can reduce the real contact area thus reduce the capillary force [34]. To be noticed, the decrease of real contact area also helps to reduce van der Waals force and is beneficial to mitigate stiction.

The fifth solution is to decrease the contact area because it also reduces the real contact area. Chow et al. propose Ball Grid Array (BGA) contact dimples, which limit the real contact area to a few tens of nanometers in diameter, exhibit acceptable contact resistance and greatly improve the reliability performance of the RF-MEMS switch [43–45].

2.1.3 Dielectric charging

The accumulation or redistribution of charge in the dielectric layer between the actuation electrodes may change the electrostatic force between the two contact surfaces. If the electrostatic force increases, the two contact surfaces will be hard to separate and even stick to each other, which is called dielectric charging induced stiction. It happens in electrostatically actuated RF-MEMS switches. The dielectric charge characterization is discussed in section 2.3.

The two different kinds of dielectric charging induced stiction are shown in the airgap-voltage (g - V) relation in Fig. 2.4. The dashed lines in the figures are the g - V curves of fresh devices. Airgap (g) is large at zero voltage; when the voltage bias (V) increases to pull-in voltage, the suspending electrode collapses and g jumps to zero; when V decreases back to pull-out voltage, the electrode releases back to its original position thus g increases again. In Fig. 2.4 (a), the net charge accumulation shifts the whole g - V curve. Stiction happens if the movable electrode cannot release when V decreases to pull-out voltage. This kind of stiction is reversible if the net charge is compensated by applying a voltage bias in the opposite polarity [34].

Another kind of stiction happens without any significant net charge accumulation. The charge is not uniformly distributed. The non-uniformity of charge can be processing related, or because of non-uniform charge injection [46]. There can be large variation of charge at local position during actuation although the net charge is zero, resulting in mechanical deformation (also observed as a narrowing of capacitance-voltage curve) [47, 48]. Stiction happens when the mechanical restoration at pull-out is prevented by the charge redistribution as in Fig. 2.4 (b). The second kind of stiction is not reversible by simply applying a voltage in the opposite polarity, because the charge variance is not a function of global charge generation [34].

Figure 2.4: Evolution of the relation between the airgap (g) and the voltage (V) because of (a) the net charge, and (b) the non-uniform distribution of charge in the dielectric layer. [34]

Here we list several methods to reduce the dielectric charging problem. The first one is to use bipolar actuation in electrostatic RF-MEMS switches. This method greatly mitigates the net charge induced shift during operation [49]. Ikehashi et al. design a charge monitor to improve the device reliability by measuring the change of pull-out voltage. The polarity of the applied voltage on the RF-MEMS switch flips when the net charge exceeds a pre-determined threshold [50]. However, this method can only prevent net charge induced stiction, but not non-uniform charge variation induced stiction.

The second method is based on the different charging rate on opposite polarities of the voltage stress. There is more than one charging mechanism [51]. The dielectric layer is directly placed on the bottom metal and is repeatably contacted by the top metal. The charge injected from the top-surface of the dielectric layer may also relate to air gap discharge and electron emission. So the charging rate relates to the actuation voltage polarity [51]. Goldsmith et al. reduce dielectric charging by using a unipolar-drive waveform of the polarity inducing less trapping than the other polarity for their dielectric material [52].

Decreasing the duration of the high voltage pulse also helps to mitigate dielectric charging. The charging is more likely to happen at high voltage. If the actuation waveform is shaped to avoid a long time of high field stress after contact, charge injection into deeper trap sites will be reduced [34]. Goldsmith et al. apply a high voltage pulse (larger than the pull-in voltage) to actuate the switch, then decrease the voltage to a medium value (larger than the pull-out voltage) to maintain the movable electrode in contact [53]. The waveform suggested by Czapslewski et al. [39] and Sumali et al. [40] have the same idea to reduce charging as Goldsmith et al., and they further

shorten the duration of the high actuation voltage to slightly longer than the switching time.

The optimization of materials and processing may also help to mitigate the charging problem. The charge trapping affinity is determined by the choice of electrode and dielectric materials, the deposition conditions, and the subsequent processes such as sacrificial layer release [34].

Many researchers also work on dielectric-less actuation switches to avoid the dielectric charging problem [22, 54, 55]. In their designs, no actuation voltage is applied over the dielectric layer of the switches, for example, by separating the RF signal and actuation electrodes [54].

2.1.4 Micro-welding

Micro-welding means the formation of a metallic bridge between the contact electrodes. For galvanic RF-MEMS switches, the Joule heating at the contact asperities can eventually melt or soften the metal and induce micro-welding, which may cause stiction [25, 56, 57]. Joule heating is also a main source of contact resistance degradation as discussed in section 2.2. To be noticed, so-called cold-welding can be induced by repeated contact operation [58]. We think the cold-welding is a kind of contact shape evolution (section 2.1.1) because it also belongs to repeated mechanical contact induced plastic deformation of the contact surfaces. We only mention the heating induced micro-welding in this section.

This failure mechanism usually happens during hot-switching. Hot-switching refers to a switch being exposed to a DC or RF signal across its contacts when it switches between up-state and contact-state [58, 59]. The small real contact area at 'a'-spots causes a significant contact resistance [36]. The high current density through these 'a'-spots is able to heat the contact metals to their melting temperature. As a result, atoms at the contact asperities may diffuse, fill the gap between the contact electrode, and finally induce welding, as shown in Fig. 2.5 [56]. If the local temperature is lower than the melting point but reaches the softening point, the adhesion force is still found to be increased [60].

To systematically study micro-welding induced failure, it is necessary to control and accelerate the formation of micro-welding for accelerated testing in a reliability study. Electrostatic discharge (ESD) or metal arcing produced between two close by contacts can easily heat the metal to the melting point and cause welding [57]. Enforced ESD can be used for accelerated testing of the micro-welding failure. Tazzoli et al. propose a method to induce micro-welding in a controllable way, which is based on ESD-like event induced by a transmission line pulser (TLP) [57, 61].

Figure 2.5: Schematic of diffusion of thermal activated metal ions near an 'a'-spot. [56]

2.2 Contact resistance degradation

Contact resistance degradation is a serious reliability problem for galvanic RF-MEMS switches, which includes a finite contact resistance change and open/short failures. As mentioned in section 2.1, both repeated mechanical contact and Joule heating can change real contact area and induce material transfer, which not only cause stiction but also cause contact resistance change. In this section, we give more details of material transfer in section 2.2.1; discuss the trade-off between adhesion and contact resistance in section 2.2.2; and introduce contamination induced contact resistance change in section 2.2.3.

2.2.1 Material transfer and evaporation

Material transfer from one contact to another has been widely observed and been considered as an essential attribute for contact resistance change [25, 56, 58–60, 62]. There are many probable mechanisms of material transfer. Besides the material transfer induced by repeated mechanical contact (section 2.1.1) and Joule heating (section 2.1.4), material transfer can also be induced by field emission, field evaporation, electromigration, micro-welding followed by Thomson effect and so on [58, 59].

Field emission and field evaporation both happen at small separations (a few Å) between contacts. For field emission, the emitted energetic electrons bombard the anode surface and may lead to heating, melting or evaporation of contact material [58]. For field evaporation, the anode metal atoms can tunnel to the cathode at an electric field lower than the threshold required for field evaporation at large separations, resulting in material transfer from

anode to cathode [58].

Electromigration happens at the small real contact area with a large current density [58]. Basically, high density conduction electrons may transfer their momentum to the atoms in the lattice of the conductor, making the metal atoms vibrate or displace. The gradual movement of metal ions in the electric field finally causes the transport of metal material, which is defined as electromigration [63]. Electromigration induces material transfer from anode to cathode.

Micro-welding makes the two contact electrodes bridge, resulting in short failure. The following Thomson effect causes the hottest point in the metal bridge between two contacts to shift towards one electrode. When the bridge breaks at the hottest point, material gets added to the electrode farther away from the hottest point [58]. The micro-welding induced short failure may also become an open failure because of the bridge breaking [56].

2.2.2 Trade-off between contact resistance and adhesion force

For a good performance of RF-MEMS switches, we expect a small contact resistance and a small adhesion force. The contact resistance decreases with increasing real contact area, whereas adhesion force increases with the real contact area [37]. Au is the preferred electrode material because of low electrical resistivity and low sensitivity to oxidation [59]. However, Au is also soft and sensitive to stiction failure. As shown in Fig. 2.6 [28], a Au-Au contact has a stable resistance but the adhesion force significantly increases during cycling which may leads to stiction failure. If a hard metal such as Ru is used, the adhesion force will remain low and a small contact resistance can also be achieved.

Hard electrodes like Ru and Pt also have a disadvantage: it is easy to have high resistance failure (probably because of contamination as discussed in section 2.2.3). The Au-Ru alloys shows much less resistance increase than pure Ru or Pt. So alloy electrode attracts people's attention. Some other studies show that the binary and ternary alloy contacts are also inert to oxidation and have an increased lifetime [64–66].

2.2.3 Contamination and frictional polymers

Contamination means that some insulator materials attach on the contact surfaces, which results in a contact resistance increase and finally leads to open failure (or high contact resistance failure). The contamination induced failure mainly occurs because of the appearance of so-called frictional poly-

Figure 2.6: The contact resistance and the adhesion force versus the number of switching cycles in case of (a) Au-Au contact and (b) Au-Ru contact. [28]

mers on the contact surfaces. Frictional polymers are organic films which develop on the contact surfaces when organic vapors or compounds are involved in the contact operation environment [58, 59]. The observed black contaminant on the contact surfaces is mainly carbon-based [29, 67].

This contamination failure mainly happens in devices using catalytically active electrodes, like Pt group metal (Pt, Pd, Ru and Rh) [58, 59]. Crossland et al. add non-catalytically active metal like Ag into Pd electrodes and significantly mitigate the problem of frictional polymerization [68]. Czaplewski et al. use less catalytically active electrodes ($\text{RuO}_2\text{-Au}$) and improve lifetime of their devices [69]. Boer et al. use contact materials with low catalytic activity and operate the switch in ultra-clean environments [70]. Their results imply that the contamination of frictional polymers is not a limiting factor in reliability of their devices any more [70, 71].

2.3 Dielectric breakdown and charging induced drift

The maximum electric field that a dielectric material can withstand under ideal conditions is called dielectric strength. Immediate dielectric breakdown happens when the electric field across the dielectric material exceeds this dielectric strength. During dielectric breakdown, a portion of the dielectric material becomes conductive. Dielectric breakdown can thus be seen as a rapid resistance degradation. In this section, we introduce the general knowledge of dielectric breakdown, dielectric leakage current, and

dielectric charging in RF-MEMS switches.

2.3.1 Introduction of dielectric breakdown

The failure mode of dielectric breakdown exists in all dielectric film involved devices. The dielectric breakdown in semiconductor devices such as MOSFET has been studied for many years and is well understood; this provides valuable basic knowledge for dielectric breakdown research in MEMS [34]. However, the application of dielectric film in MEMS devices brings more challenges and deserves more comprehensive research. One challenge is that the large lateral dimensions of MEMS (tens or hundreds of micrometers) increase the probability of defect inclusion [34]. Another challenge is that the high actuation voltage and the mechanical contact directly increase the risk of dielectric breakdown [34].

Dielectric breakdown can be mainly divided into avalanche breakdown, thermal breakdown and discharge breakdown [72]. During avalanche breakdown, the electric field accelerates electrons traveling through the insulator. The accelerated electrons will ionize atoms in the dielectric, generating an avalanche of conduction electrons in the dielectric material and causing breakdown. In other words, valence electrons get enough energy to jump to the conduction band. For thermal breakdown, the insulator is heated by the stress condition to a point where its dielectric strength drops, either by melt or by increased ionization. For example, when an applied AC stress voltage is at the dielectric's relaxation frequency, absorption of the electromagnetic energy ("dielectric loss") heats the material. Discharge breakdown can occur in gaseous spaces. In dielectric films, these can occur in pores. The occluded gas in porous material is usually ionized at a lower field strength than the solid material, and causes intermittent sparking and surface damage in this small occluded space, which accelerates the breakdown. [72]

2.3.2 Leakage current in dielectric material

Dielectric breakdown relates with the leakage current density and the Joule heating. Understanding how the charge carriers transfer in dielectric material helps to study the dielectric breakdown. The main features of several well documented sources of dielectric leakage current are listed below.

Schottky emission, also called thermionic emission, is a heat-induced flow of charge carriers. The hot electrons jump over the dielectric barrier from the metal into the conduction band of the insulator in contact with it [73]. The current is determined by the number of hot electrons which

have sufficient energy to surmount the barrier. Schottky emission is usually observed at high temperature. The barrier height is determined by the interfaces between the dielectric layer and metallic layer.

Poole-Frenkel emission (P-F emission) describes the phenomenon that the electric field causes a few electrons to hop from one place to another in the insulator [74]. The electrons can be trapped in some states called traps in the band gap of the dielectric material. Random thermal fluctuations can excite some trapped electrons to jump to the conduction band. The thermally excited electrons can move in the dielectric material for a short time, before relaxing into another trap [74, 75]. The similar emission of holes can also happen. The electrons and holes are subsequently captured again [74]. The P-F emission is temperature and electric field dependent. The hopping process happens in the dielectric bulk, thus the P-F emission is bulk controlled. It leads to a low current density because the hopping process is slow.

Fowler-Nordheim tunnelling (F-N tunnelling), also called field emission, is a quantum mechanical tunnelling process that typically occurs between a conductor and an insulator. The electrons tunnel through the triangular barrier formed by the insulator in the presence of a high electric field (E) [76]. F-N tunnelling is temperature and electric-field dependent. It is a surface controlled conduction mechanism because the triangular barrier lies at the interface between the metal and dielectric material. It is an important mechanism for thin barriers at high field, such as metal-semiconductor junctions on highly-doped semiconductors. F-N tunnelling is also widely used as the program and erase current of flash memories.

Direct tunnelling is similar to F-N tunnelling, but the electrons tunnel through the trapezoidal barrier of the dielectric at low field [77]. It has electric-field dependence and weak temperature dependence. It is strongly dependent on the thickness of dielectric film and only significantly observed in ultra-thin dielectric material (e.g. SiO₂ thinner than 2.5 nm).

Space-charge-limited conduction is a special conduction mechanism in dielectric material which happens in case of a very high injection of electrons from the electrode and few traps in the dielectric material. Space charge is a continuum of charge distributed over a space region in dielectric material. When the dielectric material and electrode has an ohmic contact, a large amount of charge carriers are emitted from the electrode and spread out in the dielectric material, producing an electrical current because of the carrier drift. The traps lower the drift mobility of carriers and thus reduce the space-charge-limited current [78]. Space-charge-limited current strongly relates to the electron velocity at the cathode and the collisions of

charge carriers in the dielectric material [79, 80]. The space-charge-limited conduction is both temperature and electric-field dependent.

2.3.3 Dielectric charge characterization and drift of device performances

At voltages below the dielectric strength, immediate dielectric breakdown is avoided, but dielectric charging between actuator electrodes still happens and is harmful to the device's reliability [12, 22–24, 34, 36, 53, 81]. It makes the pull-in and pull-out voltages drift over time, and finally induces failure such as stiction as described in section 2.1.3, premature release, or inability to properly actuate the switch [22, 34].

The dielectric charge in a MEMS device includes bulk charge, surface charge and substrate charge [34]. Charging can be detected by capacitance-voltage measurements [81, 82]. Kelvin probe force microscopy technique based on atomic force microscopy (AFM) with a conductive tip is able to draw a surface map of the charge [22, 81]. Thermally stimulated discharge/depolarization current experiments firstly induces dielectric charging, then monitors the discharge current while the temperature of the dielectric is ramped up [22]. Charging is more problematic at high temperature [22].

To describe the dielectric charging related degradation and failure, some special parameters are introduced, such as the total actuation time to failure, the shift rate of actuation voltage and the statistical variance of non-uniform charge.

The concept of total actuation time to failure is proposed by van Spengen et al. in [83]. When the RF-MEMS switch is in contact-state, charging continuously takes place under the influence of actuation voltage. It is the actuation time that determines the accumulated charging amount. The total actuation time to failure is the time it takes to attain the critical amount of charge which induces failure [83, 84]. The number of cycles to failure may differ widely depending on the actuation waveform, but the total time to failure keeps the same [83]. A simple stretched exponential for charging is used in models to predict the time to failure [85–87].

Another important parameter is the shift rate of actuation voltage, which describes how fast the pull-in voltage shifts at a certain stress condition [22]. Yuan et al. point out that the shift is linear in the accumulated charge amount [88]. Melle et al. also observe the same phenomenon and propose the lifetime can be extrapolated from the rate of shift of capacitance-voltage curve measured for a short time [89, 90]. Herfst et al. improve the method to measure the shift: the shift can be mea-

sured without fully actuating the RF-MEMS switch, but using an accurate capacitance measurement and a parabola fitting of the measured low-field capacitance-voltage curve [82, 91].

The statistical variance of non-uniform charge is an important parameter to model the non-uniform charge related phenomenon, such as narrowing of the capacitance-voltage curve (as described in section 2.1.3) [46].

2.3.4 TDDB and Weibull distribution

Besides the drift of device performances, dielectric degradation also builds up slowly until catastrophic breakdown at voltages below the dielectric strength. Time-dependent-dielectric-breakdown (TDDB) is a widely used method to investigate the dielectric breakdown at medium/low voltage stress and predict the lifetime for use conditions. Metal-insulator-metal (MIM) capacitors are usually used for TDDB measurements.

The time to dielectric breakdown not only depends on the strength of the applied medium voltage, but also relates to the measurement conditions. When TDDB is studied, a proper choice of the test conditions needs to be selected to accelerate dielectric degradation, resulting in acceptable test times but ensuring that the studied physical breakdown phenomena are representative for use conditions. Voltage and temperature are two widely used acceleration factors. Fig. 2.7 shows a typical relation between the dielectric breakdown time and the stress voltage of a MOSFET. They use voltage as the acceleration factor, measure TDDB time at around 2.5 V and predict the 10 years lifetime below 2 V.

The physical mechanisms of TDDB of many dielectric materials are not well understood yet [93]. In general, one possible mechanism is the accumulation of holes and electrons generated by impact ionization. The influence of holes is more significant because of the smaller mobility of holes. Fig. 2.8 schematically represents the widely accepted Berkeley anode hole injection model [94]. A tunneling electron is injected into the anode, and collides with another electron deep in valence band. Both electrons have final states in the conduction band, generating a hot hole which may penetrate into the oxide and create traps. The accumulation of hot holes and corresponding traps in dielectric layer may induce breakdown. If there is high accumulation of holes at the anode interface, available empty states will be within the valence band. The empty states in the valence band can be the final state of one collided electron as in Fig. 2.9 (b) or the final states of both collided electrons as in Fig. 2.9 (c). Because of these two additional minority ionization processes, a few hot holes will be left deep in the valence band Fermi-sea. These holes have large kinetic energies and

Figure 2.7: Typical relation between the TDDB time and the stress voltage of a MOSFET. In this study, the samples labeled with “SOI Uniax” show the best reliability: they can withstand the highest voltage for 10 years. [92]

Figure 2.8: A schematic anode hole injection model. [94]

significantly contribute to the breakdown.

During the TDDB measurements, the dielectric breakdown time of various devices follows a statistical distribution, instead of a fixed and predictable time. That’s because one MIM capacitor under study can be considered as lots of small parallel MIM capacitors, all being stressed and all being slightly different. The breakdown occurs when the weakest link fails. In general, if the weakest link of various devices have the same breakdown mechanism, its statistics follows the Weibull cumulative distribution, therefore the Weibull curve is widely used to analyze the TDDB behav-

Figure 2.9: Anode hot hole generation by ionization of injected electrons. (a) is the major ionization, (b) and (c) show two less probable ionizations. [95]

ior. The cumulative distribution $F(t)$ according to the Generalized Hazen Formula [96, 97] is written as

$$F(t) = \frac{i - a}{N - 2a + 1} \quad (2.1)$$

N is the number of measured devices; i is an integer satisfying $1 \leq i \leq N$; a is a parameter which is usually experimentally chosen as 0, 0.3 or 0.5 [96]. We use $a = 0.3$ for the TDDB study in this thesis.

The relation between cumulative distribution $F(t)$ and time to failure t is written in [97]

$$F(t) = 1 - e^{-(\frac{t}{\eta})^\beta} \quad (2.2)$$

β is the Weibull shape parameter, η is the Weibull scale parameter. The graph of $\ln(-\ln(1 - F(t)))$ versus $\ln(t)$ is called a Weibull-plot, where β can be easily extracted from the slope of the graph [97], as in

$$\ln(-\ln(1 - F(t))) = \beta \ln(t) - \beta \ln(\eta) \quad (2.3)$$

The β is independent of the area of capacitors under study; whereas the η relates with the capacitor area A [97, 98], following the relation

$$\frac{\eta_2}{\eta_1} = \left(\frac{A_1}{A_2}\right)^{\frac{1}{\beta}} \quad (2.4)$$

We expect a large β and η for a high quality dielectric material. A large β indicates less defects in the dielectric layer. Usually, $\beta < 1$ implies infant mortality because the failure probability density decreases with time, and the wear out of the material itself is measured in case of $\beta > 1$ [97]. The

η equals to the time when 63% of the devices break down, thus a large η indicates long expected breakdown time [97].

2.4 Reliability of PZT thin film

This section goes towards piezoelectrical RF-MEMS switches. As written in chapter 1, the reliability of PZT film is an important topic in the EPAMO project. The two main reliability questions regarding PZT actuators are: how long can the actuator remain at the on/off state under the working voltage (can it be longer than 10 years?), and how many times can the actuator switch properly? We focus on the first question in this review, by investigating the TDDB of the PZT material in a MIM structure.

The PZT TDDB models reviewed here treat the electrical domain. Although the physical nature of PZT TDDB behaviour is not clear yet, these models provide good explanation and predictions of the experiments in certain conditions.

2.4.1 Percolation model in PZT TDDB

In 1990, Jordi Sune proposed the percolation model for the description of dielectric breakdown [99]. As stress time evolves, more and more neutral traps are created in the bulk of the dielectric material. In the percolation model, it is assumed that each trap is formed independent of the others, and in a random position in the volume. A further assumption is that each trap has the same lateral extension (effective size, or trap radius). At a certain critical trap density, a type of phase change occurs (as the general percolation theory describes), which indicates that part of the dielectric material becomes conductive. At this point the probability that a string of traps connecting the two electrodes together suddenly rises from about 0 to about 1 [99, 100]. The number of traps required to make this chain is directly related to the trap radius and the dielectric thickness. In this model, a lower Weibull shape parameter β is associated with a larger trap diameter; and thinner dielectrics inherently exhibit lower β values.

A percolation model fits well the PZT TDDB measurement results according to the experiments of Chentir et al. [93]. 0.58 MV/cm constant voltage measurements at 205 °C are conducted on PZT samples of thickness from 80 nm to 400 nm, with Zr/Ti ratio of 52/48, using spin-on sol-gel processing. The top electrode is IrO₂. The Weibull shape parameter β increases from 2.21 to 3.97 when the PZT thickness increases from 80 nm to 400 nm. This is in quantitative agreement with a percolation radius of 2.5 nm.

A good percolation model should consider all elements which induce the percolation paths. Not only the traps produced by the electrical stress, but also initial defects (such as cavities) and interlayers can contribute to the percolation paths. The initial defects and interlayers make β increase slowly with the PZT thickness [93].

The source of the defects is further studied in [101]. The density and size of the cavity defects becomes larger after annealing which aims to reduce the residual stress to avoid delamination at the PZT-electrode interface. These defects are mainly located at the grain boundaries and close to the surface. The mean size of the cavities increases with the annealing temperature and time. The cavity density does not increase with the annealing time. One possible explanation is that excess lead segregates at the grain boundary, forms PbO and evaporates during annealing process, forming cavities. Because the Pb diffuses towards the surface, the cavities also concentrate to the surface. Corresponding to the percolation model, the sample with the largest cavity size has the shortest breakdown time [101].

2.4.2 High voltage and low voltage model

The PZT TDDB behaviours at high voltage and low voltage show two different failure mechanisms in ref. [102]. The observation is based on the measurements of PZT samples with 250 nm thickness and a Zr/Ti ratio of 52/48, using spin-on sol-gel processing [102, 103].

The high voltage model is used in one way to speed up the breakdown of devices: applying a voltage much larger than the working voltage. The high voltage model explains the breakdown by resistance degradation and thermal runaway. Applying a constant voltage stress from 30 to 36 V at 85 °C, the Weibull shape parameter β is 1.3, which is much lower than the β measured at low voltages [102].

The low voltage model is used for the second way to speed up the breakdown of devices: increasing the measurement temperature at the working voltage. The low voltage model explains the breakdown by percolation theory. Applying a 12 V constant voltage stress from 125 °C to 220 °C, β is 3.7 over the whole temperature range [102]. At low voltage, a resistance restoration phenomenon is clearly detected before breakdown: the current decreases with time. It can be attributed to the trapping effect. Ionized Pb vacancies can be trapping centers, and neutralize or reorganize the space charge [103, 104].

The authors extrapolate the experimental results under voltage stress from 10 to 14 V from 180 °C ~ 220 °C to 85 °C. The measured time to breakdown can be fitted well by the Arrhenius law [102]. Not much study

of Arrhenius law in PZT reliability is found, and it deserves more research. This thesis also treats the temperature extrapolation in chapter 7. The extrapolation result of the low voltage model fits both E model and $\sqrt{E}$ model well in [102].

Those TDDB experiments indicate various competing degradation processes in PZT thin film. The high voltage model and low voltage model represent different failure mechanisms. Temperature acceleration method may be a good way for PZT TDDB study [105]. More research is necessary to understand the physics of PZT TDDB, which is the aim of Chapter 6 and 7 of this thesis.

2.4.3 Infant mortality model

The TDDB research on multilayer ceramic actuators from various groups show a Weibull shape parameter smaller than 1 [106,107], which indicates an infant mortality [97]. J. S. Lee et al. [106] propose a so-called infant mortality model to describe the intrinsic defects induced breakdown. It is built for the multilayer PZT actuators, using $0.2\text{Pb}(\text{Mg}_{1/3}\text{Nb}_{2/3})\text{O}_3$ - $0.8\text{Pb}(\text{Zr}_{0.475}\text{Ti}_{0.525})\text{O}_3$ (PMNZT) ceramic powders synthesized by calcination process and Ag-based internal electrodes. Samples are measured under 2 kV/mm AC electric field at 50 °C in 30% relative humidity¹. The leakage current is monitored to determine the number of cycles to breakdown. The Weibull shape parameters β of all samples are smaller than 1 (0.43 to 0.59).

The reported reason of this breakdown is the processing defects and handling failures [106]. The defects are most possibly created in the cofiring process, which is needed to reduce the delamination phenomenon in the PZT layer. The processing defects are generated because of the residual stress between the PMNZT ceramic and the electrode during the cofiring process [106]. It has been widely accepted that the residual stress between the piezoelectrically active and passive regions is a major degradation mechanism of multi-layer actuator [108–110]. The high temperature cofiring process induces the residual stress because of the large mismatch in thermal expansion coefficient between electrode and PZT layers during the cofiring and subsequent cooling process.

The electrode material influences the reliability of the stack. The reliability is improved by adding small amount of PMNZT ceramic powders into the Ag electrode. Adding low concentration of PMNZT ceramic in

¹As discussed in chapter 7, the humidity of measurement environment greatly influences the PZT reliability. We should be careful with the interpretation of measurements in humid environments

Figure 2.10: The fluctuating and gradually increasing I - t curve of a PZT capacitor measured with several times of suspension of the constant voltage stress. Measurement details refer to expression of Fig. 7.6. This I - t curve is similar to the I - t curve published in [111].

Ag electrode can reduce the mismatch in the thermal shrinkage, resulting in an increased reliability [106]. However, a high concentration of ceramic makes the reliability worse. This is because the skeleton structure of the Ag electrode leads to a decrease of the effective electrode area and causes a local non-uniform electric field at the metal-dielectric interface [106].

2.4.4 Electromigration model

In the context of PZT degradation, electromigration refers to the process that the electrode atoms at anode are ionized and drift to cathode, creating a conductive pathway in the dielectric layer and hence causing breakdown. The continuous local breakdown causes further damage to the structure: cracks and fused material [111–113].

This model is built from the high humidity measurements. The fluctuating I - t curve gradually increases even with suspension of the constant voltage stress. We measure I - t curves similar to the I - t curve described in [111], as shown in Fig. 2.10. This fluctuating, increasing current is accompanied by an increasing of visible broken spots on the cathode [111].

Here is a proposed explanation of the fluctuating, increasing I - t curve. The metallic ions of anode electrode migrate to the cathode under a DC bias stress. It induces electrochemical reduction and creates a metallic filament on the cathode. The metallic filament grows towards the anode. When the distance between the filament tip and the anode is short enough, an electric

breakdown occurs and induces a transient current. The electric breakdown may partly destroy the filament and clear the short circuit path, resulting in a decrease of current. The metallic filament quickly regrows and causes successive breakdowns at the same site. The increase of the overall leakage current is because of the development of the localized damage sites. [111]

2.5 Other failure modes in RF-MEMS switches

Besides the failure modes discussed above, some other failure modes are also important in RF-MEMS switches. In this section, we briefly introduce creep and RF power related reliability issues.

2.5.1 Creep

Creep is the time-dependent relaxation of the movable electrode under mechanical stresses [22, 36]. When a voltage stress is applied between the movable electrode and the fixed electrode at the up-state, the suspended electrode tends to bend. This results in an increase of capacitance. If the stress is not strong enough for a plastic deformation, the creep is reversible.

Creep is hardly observed when the temperature is lower than 0.3 times of the melting temperature of the electrode material [22]. The presence of immobile and hard precipitates in the suspended electrode is able to inhibit creep [22]. The insulator can be used as a creep-resistant structural layer as part of the movable electrode [22].

2.5.2 RF signal related reliability issues: power handling and heating

Power handling capability describes the maximum RF power that the RF-MEMS switches can handle to ensure a proper operation [1]. The RF voltage induces a force on the suspended electrode when the actuation voltage (DC bias) is zero. If the RF power is large enough, this force can induce self-actuation, which actuates the switch to contact-state [1]. When a certain actuation voltage is applied, the RF voltage exerts an extra force to the movable electrode, which can act as holding voltage or even actuate the switch [22, 114].

RF power also influences the device reliability by heating. The RF power tends to initiate higher current in the outer features of the device which is called skin effect [115]. This may lead to localized hot spots and local melting [22].

2.6 Conclusion

This chapter gives an introduction of reliability problems of RF-MEMS switches.

The main failure modes of RF-MEMS switches, namely stiction, contact resistance degradation, dielectric charging and breakdown, are discussed in detail. Both stiction and contact resistance degradation can be caused by either a mechanical reason (repeated contact) or an electrical reason (Joule heating, field emission, field evaporation and so on). The trade-off between adhesion force and contact resistance should be considered to choose a suitable material of contact electrodes. The capillary effect and dielectric charging also contribute to stiction problem. Surface contamination caused by frictional polymers can increase the contact resistance and finally lead to open failure. The failure mode of dielectric charging and breakdown has been widely studied in semiconductor devices; and the large dimensions and high actuation voltages of MEMS devices bring new reliability questions in this failure mode. The dielectric breakdown mechanisms, the origin of leakage current, charging and TDDB mechanisms are introduced in this chapter. The TDDB models of PZT thin films (treated in the electrical domain) are summarized, including the percolation model, the high voltage and low voltage models, the infant mortality model and the electromigration model. These models provide good explanations of the experiments in certain conditions, although the physical nature of PZT TDDB behavior is not clear yet. Finally, creep and RF power related reliability issues are briefly introduced.

Chapter 3

Capacitance measurement of RF-MEMS switches

RF-MEMS capacitive switches have great chance to penetrate the high-volume handset market [22]. As reviewed in chapter 2, these devices combine a high tuning ratio with a good quality factor, making them an attractive choice for tunable impedance matching at the antenna. For process development and process control, electrical testing at wafer level of these components is essential. A delicate calibration and de-embedding enables an accurate capacitance measurement, which is the basics of characterization and reliability studies of the capacitive RF-MEMS switches.

Chapter 3 focuses on the comparison of various measurement methods on the device under test (DUT): the capacitive RF-MEMS switches. In section 3.1, the origin of short and open parasitics are explained in detail. Five capacitance measurement methods are introduced, managing to measure the small-signal capacitance across 11 orders of magnitude in frequency. The capacitance-frequency (C - f) curve of the DUT is shown to reflect the quality of calibration and de-embedding. In section 3.2, capacitance-voltage (C - V) curves measured by the five approaches mentioned in section 3.1 are compared; some abnormal phenomena, such as frequency dependent changes of C - V curves and intermediate state, are detected and discussed.

3.1 Calibration, de-embedding and measurement principles

The wafer level measurement of capacitive RF-MEMS switches is complicated because the devices are very sensitive to the ambient before packaging, and the devices have relatively low capacitance of only a few pico-

Figure 3.1: Top view microscope image of the RF-MEMS capacitive switch under study. The A - A' and B - B' lines indicate the position of the cross-sections in figure 3.2 and figure 3.3.

farads which is comparable with the parasitic capacitance. It is important to identify where are the parasitics and eliminate their influence in the measurement of capacitance of device under test (C_{DUT}).

3.1.1 Device and system parasitics

The studied RF-MEMS capacitive switches are fabricated on a Si substrate. Fig. 3.1 shows the top view of the DUT designed for ground-signal-ground (GSG) probing. The large perforated square ($300 \times 300 \mu\text{m}^2$) is the movable top electrode, which is connected to four anchors by the curved springs and the two ground (G) pads. The underlying bottom electrode is connected to the signal (S) pad. All metallization is aluminum. A $\text{SiO}_x/\text{SiN}_x$ dielectric layer is deposited on the bottom electrode.

The schematic cross sectional view of the bond area (A - A') is shown in Fig. 3.2 (a). Parasitic capacitance exists between the bond pads, through air (C_{GS}) and through the silicon substrate (C_G underneath the ground pads and C_S underneath the signal pad). Parasitic resistance exists between the bond pads through the silicon substrate (R_{BP}). Given that the outer two pads are both grounded, the equivalent circuit in the bond pad area A - A' can be simplified as shown in Fig. 3.2 (b). C_{BP} and R_{BP} are the equivalent parasitic capacitance and resistance through the silicon substrate respectively.

The schematic cross sectional view of the DUT at B - B' is shown in Fig. 3.3 (a). Besides C_{DUT} , parasitic capacitance is found around the top electrode via the substrate (C_{BE} underneath bottom electrode and C_{G+}

Figure 3.3: (a) The schematic cross sectional view of DUT from B - B' in the top view image, (b) the small-signal equivalent circuit of the parasitics around the movable top electrode.

Figure 3.4: The small-signal equivalent circuit of all parasitics of the device and measurement system.

than $R_{BP}C_{BP}$. The compensation of parasitics around bottom electrode and springs should be carefully considered in the design of open reference structure. In addition, the length of connection lines on the wafer greatly influences the short parasitics and should be considered in the short compensation. Little literature is found on this.

3.1.2 Various capacitance measurement techniques

Five C - V measurement techniques will be used in this study. Section 3.1.2 will introduce how various techniques measure capacitance and how to compensate the parasitics mentioned in section 3.1.1.

3.1.2.1 Measurement principles and experiments

The five capacitance measurement techniques used in this study are quasi-static (QS) measurement, low-frequency (LF) measurement (1 mHz to 10 Hz), two kinds of high-frequency (HF) measurement (10 kHz to 1 MHz), and radio-frequency (RF) measurement (100 MHz to 1 GHz). All techniques aim to qualify the small-signal capacitance of a DUT, that is, $C = \frac{\partial Q}{\partial V}$.

A Keithley 595 quasi-static meter is used for the QS measurements. A feedback charge method is used by this quasi-static meter, as shown in Fig. 3.5. C_F is a constant feedback capacitor. The input voltage (V_{in}) applied to the signal pad is a voltage stair, the ramp of which is determined by two parameters: delay time and voltage step (V_{step}). By measuring the output voltage (V_{out}), we get the charge on the feedback capacitor, which is $Q_F = V_{out} \times C_F$. The change of charge (ΔQ) on the capacitance being measured (C_m) equals to that on C_F . By measuring V_{out} at consecutive input voltage steps and calculating the change of charge, the measured capacitance can be obtained by $C_m = \frac{\Delta Q}{V_{step}} = \frac{Q_{F2} - Q_{F1}}{V_{step}} = \frac{(V_{out2} - V_{out1}) \times C_F}{V_{step}}$ [116].

Both an HP 4284A LCR meter (LCR) and a Keithley SCS 4200 (K4200) are used for the HF measurements. For this technique, a DC voltage bias and a small-signal AC voltage (V_{AC}) are both applied to the signal pad. The corresponding AC current (I_{AC}) and the phase difference between I_{AC} and V_{AC} are both measured, thus the impedance of the device being measured can be calculated by $Z_m = \frac{V_{AC}}{I_{AC}}$. Normally, the measured capacitance is calculated from the impedance by $C_m = \frac{1}{\omega \text{img}(Z_m)}$, ω is the angular frequency of the applied AC voltage, $\text{img}(Z_m)$ is the imaginary part of the measured impedance.

Figure 3.5: Schematic circuit of feedback charge method.

The LF measurements are conducted using the Keithley SCS 4200 with two pre-amplifiers (Model 4200-PA) installed on the back panel of the Keithley SCS 4200 [117]. In the set-up of the LF measurements, starting values for measured capacitance and resistance should be given. Only with well-chosen values (from previous measurements or calculated estimates) we obtain good convergence towards the expected capacitance. The LF measurement technique is recently introduced by Keithley and we are one of the first groups to experiment with it.

A ZVB20 network analyzer [118] is used for 1-port RF C - V measurements. Open/short/load calibration is carried out on a standard calibration pattern. The device-related parasitics are consecutively eliminated by de-embedding (as described later). The measured device impedance is calculated from S_{11} by the equation $Z_m = 50\Omega \times \frac{1+S_{11}}{1-S_{11}}$ [119]. Again, the measured capacitance can be obtained from the imaginary part of the measured impedance by $C_m = \frac{1}{\omega \text{img}(Z_m)}$.

All the measurements are conducted at room temperature. The devices are not encapsulated. Hence a preparation process is needed to remove the moisture: a 10 minute dry air flush at room temperature followed by a 10 minutes heating step in dry air at 150 °C. The DUT (or the open/short reference structures) are contacted with two micro-manipulators, with the exception of the RF measurement where a GSG probe is used. A DC voltage sweep with step of 0.1 V and a AC voltage of 80 mV are used for all the LF and HF C - V measurements in this chapter. For RF C - V measurements, power of 0 dB is used, thus the AC voltage is approximately 223 mV according to Table 3.1. The network analyzer has a standard input/output impedance (Z), so the conversion between the power (P) and the AC voltage (V_{AC}) can be calculated by $V_{AC} = \sqrt{P \times Z}$ as shown in Table 3.1.

Power	Power (P)	Impedance (Z)	AC voltage ($\sqrt{PZ}$)
-20 dBm	10 μ W	50 Ω	22.3 mV
-10 dBm	100 μ W	50 Ω	70.7 mV
0 dBm	1 mW	50 Ω	223 mV
10 dBm	10 mW	50 Ω	707 mV
20 dBm	100 mW	50 Ω	2.23 V
30 dBm	1 W	50 Ω	7.07 V

Table 3.1: Conversion between the power and the AC voltage of the network analyzer ZVB20.

3.1.2.2 Measurement of open and short compensation

We expect negligible influence of the DC bias on the parasitic elements, and characterize open and short reference structures at 0 V bias.

One DUT is sacrificed to make an open reference structure: the top electrode is mechanically removed from the device, and the damaged structure is then used as the open reference structure, as shown in Fig. 3.6 (a). This approach does disconnect the two ground pads, so we divide the obtained open impedance by two when the devices are probed with two micromanipulators (for QS, LF and HF C - V measurements). For RF C - V measurements, the two ground pads are connected through the GSG probe, so the obtained “open” impedance is correct. The open-correction is purely capacitive at our measurement frequency range from quasi-static to 1 GHz.

The short reference structure is shown in Fig. 3.6 (b). The short impedance is more than 6 orders of magnitude lower than the open impedance and DUT impedance at frequencies below 10 MHz, so the inductive short correction is only important in our RF measurements. There is a significant difference between the parasitic short inductance of DUT and that of the provided short reference structure on the wafer. As estimated in Fig. 3.6 (b) and (c), the induction-loop area of DUT is smaller than that of the short reference structure. L_s as obtained from the short measurement can be corrected manually by multiplying L_s with the loop area ratio of 160/270.

The measured short and open compensation impedance (Z_s and Z_o) connect to the DUT as in the schematic circuit of Fig. 3.7. To be noticed,

Figure 3.6: (a) Open reference structure, (b) short reference structure, (c) device under test.

Figure 3.7: Schematic circuit of open, short and DUT connection.

Z_{mDUT} is the measured impedance of DUT and Z_{DUT} is the real impedance of DUT. Therefore, $Z_{DUT} = \frac{Z_o \times (Z_{mDUT} - Z_s)}{Z_o - Z_{mDUT} + Z_s}$.

3.1.3 Capacitance/impedance-frequency analysis

After a careful measurement of DUT, we would like to check whether the calibration and de-embedding is good, by plotting the capacitance-frequency curve of the DUT. In addition, the value of each element in the equivalent open/short parasitic circuits will be obtained through data fitting of impedance-frequency curves of open/short reference structures.

3.1.3.1 Capacitance-frequency curves of DUT

The capacitance-frequency ($C-f$) curve of a DUT should be a straight line if all the parasitics are calibrated and de-embedded well in the frequency range of our measurements (except around resonant frequencies, because of

the mechanical effects which will be discussed in chapter 4). To check it, exemplary C - f curves of DUT, obtained by the five measurement techniques, are shown in Fig. 3.8. In these C - f curves, the obtained capacitance values in the up-state (small capacitance) and down-state (large capacitance) of the switch are shown from all five measurement approaches. As the QS measurement does not associate with a measurement frequency, the QS capacitance values are indicated with dashed lines. HF measurements (using LCR and K4200) below 10 kHz are inaccurate, as the device capacitance is too low, hence these data are not shown here.

When the short compensation is set to zero in calculation, the LF and HF C - f curves are not influenced much but the RF C - f curve goes up with frequency, indicating an non-compensated inductive parasitic, as in Fig. 3.8 (a). If the measured impedance of the short reference structure is used in the calculation without correction, as in Fig. 3.8 (b), the RF C - f curve is over-compensated. When a corrected short compensation, as pointed out in Fig. 3.6 (b) and (c), is used, we get a straight C - f curve as in Fig. 3.8 (c). The obtained average value of C_{DUT} is 5.65 pF and 0.3 pF at down-state of 15 V and up-state, respectively, almost independent of frequency. This indicates an accurate calibration and de-embedding.

3.1.3.2 Impedance-frequency curves of open/short reference structures

From the impedance-frequency (Z - f) curve of short/open reference structure, it is possible to do data fitting and determine the value of each element (resistance, capacitance and inductance) of the equivalent circuit of short/open parasitics. The equivalent parasitic circuit together with DUT (a pure capacitance at either up-state or down-state) can be used as a compact model of the RF-MEMS switch. A simple and accurate equivalent parasitic circuit will be helpful to the application of the RF-MEMS switch, such as the RF-MEMS switch involved circuit or system design. All the parasitics are involved in the circuit of Fig. 3.4, however, this parasitic circuit is too complicated for data fitting. Therefore, we try the parasitic circuit from simple to comprehensive. After comparing the data fitting accuracy, we finally choose the simplified short and open parasitic circuit as in Fig. 3.9 (a) and (b). C_{sub} and R_{sub} are all the parasitic capacitance and resistance through the substrate; C_{air} is the parasitic capacitance through the air; L_s and R_s are the inductance and resistance caused by the connection lines on the wafer. R_s is not shown in Fig. 3.9 (c) because it is negligible compared with R_{sub} . C_{DUT} connects in parallel with open parasitic circuit, so Fig. 3.9 (c) is used as the model of the RF-MEMS switch.

(a)

(b)

(c)

Figure 3.8: Capacitance-frequency curves of DUT at both up-state (small capacitance) and down-state (large capacitance) by using (a) zero short compensation, (b) all the measured short compensation, and (c) manually corrected short compensation according to the inductance-loop area ratio.

Figure 3.9: (a) short and (b) open equivalent parasitic circuit; (c) MEMS switch

The weighted least square fitting method is used to calculate the value of each element in the open and short equivalent parasitic circuits. The fit values are: $L_s = 2.73 \mu\text{H}$; $R_s = 1 \Omega$; $C_{sub} = 1 \text{ fF}$; $R_{sub} = 10 \text{ M}\Omega$; $C_{air} = 0.09 \text{ pF}$. The data fitting results of up-state and down-state MEMS switch measured by LF and HF C - V techniques are shown in Fig. 3.10. The y-axis in Fig. 3.10 is the absolute value of the complex impedance. We calculate the accuracy of the measured impedance and get its error bar. The error bar of the up-state impedance from 10 kHz to 100 kHz is large, because the impedance of the DUT in this frequency range is close to or even larger than the upper limit of the AC impedance which can be accurately measured by the instrument. The solid lines are the fit results. At up-state at the DC bias of 1 V, $C_{DUT} = 0.26 \text{ pF}$; at down-state at DC bias of 26 V, $C_{DUT} = 6.14 \text{ pF}$. The C_{DUT} obtained from the data fitting matches the C_{DUT} directly obtained from the C - f curve.

3.2 Capacitance-voltage measurements of RF-MEMS switches

An accurate C_{DUT} can be measured after the calibration and de-embedding as described in section 3.1. By comparing the C - V curves of DUT measured by five various techniques, we have been able to develop a comprehensive understanding of the RF-MEMS switches.

Figure 3.10: Fitting results of DUT at up-state and down-state.

3.2.1 C - V comparison measurement by five approaches

3.2.1.1 Work principle of RF-MEMS switches

To analyze the C - V curves of DUT, we would like to firstly introduce the basic work principle of DUT: the electrostatic capacitive MEMS switch. The DUT has two states, which are up-state (small capacitance) and down-state (large capacitance), as shown in Fig. 3.11 (a). When the potential difference between two electrodes increases to pull-in voltage (V_{pi}), the electrostatic force overcomes the spring force and pulls the top electrode down, therefore, the DUT switches to down-state. Oppositely, when the potential difference between two electrodes decreases to pull-out voltage (V_{po} , also called release voltage), the electrostatic force cannot hold the top electrode any more and releases it, thus the DUT switches to up-state. There are both positive and negative V_{pi} and V_{po} as shown in the classical small-signal C - V curve measured at 800 kHz in Fig. 3.11 (b). In the following measurements, we will see the small-signal C - V curve deviates from the “classical behavior” at low frequency.

3.2.1.2 Relation between capacitance of DUT and DC bias measured by five approaches

The de-embedded full C - V curves of one DUT measured by these five approaches are shown in Fig. 3.12. The DC bias is swept from -25 V to

Figure 3.11: (a) Schematic two states of an electrostatic capacitive RF-MEMS switch, (b) typical C - V curve measured at 800 kHz.

25 V, so the DUT switches to up-state at negative V_{po} and switches back to down-state at positive V_{pi} . The DC bias of QS and LF C - V measurements are limited to 20 V. Upon closer inspection, the C - V curves measured by different instruments exhibit some differences, in particular near V_{pi} and V_{po} . When the DC bias is approaching V_{po} , the HF (and RF) capacitance gradually goes down on one hand, and LF (and QS) capacitance remains constant then goes up on the other hand. When the DC bias is approaching V_{pi} , the HF (and RF) capacitance slightly increases on one hand, and LF (and QS) capacitance significantly increases on the other hand. A similar phenomenon is observed when the DC bias is swept from +25 V to -25 V. The AC voltage influences the pull-in voltage in LF measurements (as discussed later in this chapter) but does not influence the shape of the C - V curve.

A more detailed comparison of the C - V curves at down-state near V_{po} is shown in Fig. 3.13. In the LF C - V measurements (Fig. 3.13 (a)), C_{DUT} decreases with a rate of 0.02 pF/V from -20 V to -11 V (independent of the

Figure 3.12: Comparison of de-embedded C - V curves of an RF MEMS capacitive switch measured by five approaches.

frequency). Between -11 V and V_{po} , C_{DUT} keeps constant or even increases with DC bias voltage. The QS C - V measurement (Fig. 3.13 (b)) is similar, be it somewhat obscured by the lower measurement accuracy. In addition, spikes appear at V_{po} of both LF and QS C - V curves. In RF, K4200 and LCR measurements, C_{DUT} decreases with an average rate of 0.025 pF/V between -20 V and -11 V. The decrease is somewhat faster between -11 V and C_{DUT} (see Fig. 3.13 (c), (d) and (e)).

Further, we find that near V_{po} the down-state LF C - V curve varies a lot from device to device as shown in Fig. 3.14 (a), whereas all other measurements at down-state reproduce very well. In some cases, the down-state capacitance increases significantly when the DC bias approaches the pull-out voltage. At up-state, a measurement on six different devices on the same wafer, with the same design, gives very comparable results as in Fig. 3.14 (b). We will discuss this difference in chapter 4.

To check whether the LF/QS capacitance deviates from HF/RF capacitance when the DC bias is much larger than V_{po} , we manually increase the limited DC bias of LF C - V measurement, by adding an external DC bias of 15 V to the virtual ground of the LF C - V measurement in K4200. The de-embedded C - V curves of DUT are shown in Fig. 3.15. The DC bias is swept from 0 V to -30 V, then is swept back to 0 V (the sweep direction is indicated with arrows). The spike at -15 V is an artefact caused by the

Figure 3.13: Comparison of the capacitance of DUT at down-state: (a) LF, (b) QS, (c) HF: K4200, (d) HF: LCR and (e) RF $C-V$ measurements.

Figure 3.14: (a) down-state LF $C-V$ curves near V_{po} , measured on 6 equally designed devices on the same wafer; and (b) up-state LF $C-V$ curves near V_{pi} from the same 6 devices.

external DC bias as marked with the middle oval in Fig. 3.15. At high DC bias, there is no difference among capacitances measured at any frequency. The discrepancy between LF and QS capacitances on one hand, and HF and RF capacitance on the other, only happen near V_{pi} and V_{po} as marked with rectangular boxes in Fig. 3.15. At positive bias, the device exhibits the same behavior.

The classical $C-V$ curve, in which the capacitance gradually decreases with reducing electric field at downstate, is commonly and consistently observed on RF MEMS capacitive switches. It is attributed to the flattening of the dielectric-metal interface or zipping effects of the top electrode membrane and the springs [1, 2, 120–123]. The HF and RF $C-V$ curves fit this

Figure 3.15: Comparison of de-embedded quasi-static, low frequency, high frequency and radio frequency C - V curves of an RF MEMS capacitive switch.

common understanding. However, in the QS and LF C - V measurements, the C_{DUT} at down-state does not decrease with the absolute value of DC bias voltage close to the pull-out voltage. Instead, C_{DUT} increases when the DC bias is close to V_{po} , so either the measured values are incorrect, or the commonly accepted contact model is at least incomplete.

It should be noted that the QS C - V measurement is error-prone upon the sudden switching of the device, where the assumption of quasi-static conditions does not hold. (At pull-out, an infinitesimal change in DC voltage leads to a large capacitance change.) Therefore, we think the spikes at V_{po} and V_{pi} of QS C - V curves are artefacts. In addition, similar but smaller artefacts also randomly appear at V_{po} and V_{pi} of LF C - V curves as shown in both Fig. 3.12 and Fig. 3.15. Further explanation will be given in chapter 4.

To verify if the observed differences between QS/LF and HF/RF C - V curves are caused by the chosen experimental conditions, a range of sanity checks is performed. We find the same C - V behavior on various switches, with various measurement set-ups, and under a wide variety of measurement settings. Although the value of C_{DUT} and pull-in/pull-out voltages may slightly shift from device to device, the LF C - V curve exhibits a similar shape with an applied frequency from 0.01 Hz to 10 Hz (Fig. 3.13 (a)) and with an applied rms AC voltage from 20 mV to 500 mV (Fig.

Figure 3.16: The LF C - V curves measured with various AC voltages.

3.16). (This indicates that our small-signal assumption is valid; it should be noted that large signal biasing of this device would require a separate treatment.) It further overlaps with QS C - V measurements, independent of the QS sweep rate (Fig. 3.13 (b)).

The LF and QS measurements typically have a slower voltage sweep (the sweep rate can be down to 0.05 V/s as shown in Fig. 3.13 (b)) than the HF methods. We investigate whether the different measurement time scale is the cause for this effect, by artificially slowing down the HF C - V sweeps. The measurement time does not influence the C - V curve in these HF measurements even with 4 seconds hold time and 4 seconds wait time, excluding relaxation effects as a possible explanation of the differences.

Environmental humidity does not affect the measured C - V curves. Further, no measurement-induced device changes have been observed on any of the measurement techniques or devices. From our experiments reported above, the observed frequency dependence of the small-signal capacitances stems from the device itself. The origin of this discrepancy will be explained in chapter 4.

3.2.2 Intermediate capacitance state

Besides the frequency dependent discrepancy of C - V curves, we sometimes detect another abnormal behavior in the C - V curves: an ‘intermediate state’ between the down-state and the up-state (cf. Fig. 3.12, Fig. 3.15

Figure 3.17: C - V curves measured by five techniques, around the intermediate state observed at pull-out.

and Fig. 3.16). Fig. 3.17 zooms in on the intermediate state of the C - V curve of Fig. 3.12. The capacitance of this state is 1.5 pF in K4200, LCR and RF C - V measurements and 2 pF according to QS and LF C - V measurements.

To further investigate this intermediate state, the topography of the device is measured by optical means using an MSA-400 Micro-system-analyzer. Fig. 3.18 (a) shows the topography of the device at down-state. The red line drawn in the figure runs from the center of the movable top electrode to the fixed end of one anchor. We quantify the height of the device along this line as a function of the applied DC bias. As shown in Fig. 3.18 (b), the center of the top electrode moves up slightly before the whole electrode comes up, likely corresponding to a lower capacitance than the full down-state. Therefore, one possible explanation of this behavior is that the top electrode is not flat. The center part is higher than the edge part as a result of the fabrication process. A second factor may be that the electric field at the edge of the electrode is larger than in the center, so the electric force at the edge is larger.

According to the measurement results of more than 10 devices, the intermediate capacitance is always there but differs somewhat from device to device. It also changes slightly when we repeat the voltage-sweep measurement. The different capacitance values observed in the intermediate state among different measurements may then be because of different area of the center of the top electrode which slightly moves (see Fig. 3.18 (b)).

Figure 3.18: Dynamic topography of the device, (a) the device in down-state, (b) the height of the surface of the device along the red line in (a) under different DC bias voltages.

3.2.3 Pull-in voltages

The V_{pi} values of one DUT measured by the different techniques do not coincide exactly, as shown in Fig. 3.19. For the RF C - V measurement, V_{AC} is changed through changing the power of the network analyzer, as converted in Table 3.1. In QS C - V , V_{AC} is not determined, and consecutive measurements lead to values randomly differing up to 0.2 V. Hence V_{pi} (18.5 V to 18.7 V) from the QS C - V curve is not shown in this graph. The shift of pull-in voltage in publications is measured at a constant AC voltage and is usually considered a degradation of the switch, which mainly relates to the charging [81], and sometimes relates to thermal expansion [124]. Because the results in Fig. 3.19 can be repeated, we think the shift of pull-in voltage is caused by the change of AC voltage instead of the degradation

Figure 3.19: Pull-in voltages measured by four methods, as a function of the applied AC test signal amplitude.

of the device.

The V_{pi} obtained from LF C - V measurements, where the measurement frequency is much lower than the device switching speed, has almost linear dependence on applied AC voltage. When the test signal has a larger amplitude, the instantaneous electric field may be significantly higher than the DC bias alone, which can lead to earlier pull-in. Fig. 3.16 also clearly shows this earlier pull-in when a larger AC voltage is used. The V_{pi} observed in HF C - V curves is only slightly influenced by AC voltages. We expect the same V_{pi} measured by various HF C - V equipments. The difference between V_{pi} measured by K4200 and LCR may be induced by the inaccuracy of DC bias in various equipments. In RF C - V measurements, the AC voltage hardly affects V_{pi} . One possible reason is that the mechanical displacement of the top electrode cannot react fast thus it hardly displaces with the AC voltage at HF or RF. Therefore, a higher AC voltage will not lead to earlier pull-in at HF or RF. More detail of the mechanical reaction will be discussed in chapter 4.

3.3 Conclusion

The small-signal capacitance of capacitive RF-MEMS switches is measured across 11 orders of magnitude from 10 mHz to 1 GHz. An equivalent circuit, representing the origin of the parasitics, is given. After calibration and de-embedding, the device capacitance is almost frequency independent. Using weighted least square fitting, the values of each element in simplified open

and short parasitic circuits are calculated.

The C - V curves measured by five different methods are compared. The LF and QS C - V curves yield different results from classical HF and RF C - V curves. The LF and QS capacitance is up to 200% and 25% higher than the HF and RF capacitance around the pull-in and pull-out voltages. The root cause of this difference will be explained in the following chapter. An intermediate capacitance state is observed at pull-out and associated with the stepwise raising of the moving electrode. The relation between V_{pi} and V_{AC} varies from LF C - V technique to HF/RF C - V technique. It may relate with the mechanical characteristics of the top electrode displacement.

Chapter 4

Transducer model of RF-MEMS switches

RF-MEMS capacitive switches are two-terminal devices, with a small-signal capacitance determined by the applied DC bias voltage [1, 2, 125]. The capacitance-voltage (C - V) measurement is the most commonly used technique to obtain the main electrical parameters of RF-MEMS switches [121, 125]. In the scientific literature on these devices, all C - V measurements reported to date are conducted at frequencies above 1 kHz [121, 126–129]. However, a comprehensive understanding of the devices from DC to RF is necessary for accurate modeling in applications, because this type of switch is often used to process a signal with a very broad spectrum, e.g. in the front end of a wireless receiver.

It has been described in chapter 3 that significant deviations from the classical C - V curves are observed when the small-signal capacitance of devices under test (DUT), the RF-MEMS capacitive switches, are measured at low-frequency (LF) or measured by quasi-static (QS) technique. The classical C - V curves are obtained in the measurement at high-frequency (HF) and radio-frequency (RF). Chapter 4 brings new experimental results and an explanation for the differences between the LF/QS capacitance and HF/RF capacitance of DUT. In section 4.1, a one-dimensional transducer model is introduced to explain the frequency dependent discrepancy in C - V curves at up-state. Section 4.2 focuses on the down-state model. Both the mechanical displacement of the top electrode under various conditions and the C - V curves in vacuum indicate that the squeeze film effect plays an important role in the HF capacitance measurement.

Figure 4.1: Schematic cross sectional view of DUT of top electrode area with dimensions used in the transducer model.

4.1 Transducer model of the up-state capacitance

We firstly focus on the up-state C - V curves. The reason why the LF/QS C - V curves deviate from the classical HF/RF C - V curves relates with the mechanical characteristics of the DUT. A transducer model is introduced in this section to represent the coupling between mechanical and electrical characteristics of the DUT.

4.1.1 Application of the transducer theory to the up-state capacitance

The DUT can be considered a voltage controlled electrostatic transducer. Under voltage control, the combination of voltage source and the switch may try to achieve minimum free internal energy (U), i.e. the sum of capacitive and mechanical (spring) energy.

For the up-state, one can derive an equation to describe the impedance of the switch, assuming a one-dimensional system with a mass-loaded spring with spring constant K , moving electrode mass m , and damping characterized by a viscous damping factor α . The schematic cross sectional view of DUT of top electrode area is shown in Fig. 4.1. The position of the upper surface of the bottom electrode is defined as height 0; the original height of the top electrode x_0 is 3 μm ; the thickness of the dielectric layer x_d is 560 nm; x is the DC bias dependent height of the top electrode. The roughness of under-surface of the top electrode σ is also mentioned in Fig. 4.1 because it is important for the down-state model as discussed in section

4.2. According to the study of H. M. R. Suy et al. on the same samples [26], the surface height distribution of top electrode is close to Gaussian distribution (as drawn in Fig. 4.1), which has a standard deviation of σ and a midplane (zero point) at the position of $x = 3\sigma + x_d$. Our AFM measurement indicates that the roughness of the dielectric layer is too small to be considered in the contact model, which is also corresponding to the study of H. M. R. Suy et al. in [26]. The parallel-plate capacitance is given by

$$C_p = \frac{\varepsilon_0 A_p}{x_e} \quad (4.1)$$

With x_e the equivalent distance of the top electrode to the bottom electrode (see also Fig. 4.1):

$$x_e = x - x_d + \frac{x_d}{\varepsilon_r} \quad (4.2)$$

In these equations, A_p is the parallel-plate capacitor area; ε_0 the permittivity in vacuum, and ε_r the relative permittivity of the $\text{SiN}_x/\text{SiO}_x$ dielectric layer on the bottom electrode.

The internal energy of the transducer and its partial differential are given by:

$$U = \frac{V_{\text{DC}}^2 \varepsilon_0 A_p}{2x_e} + \frac{1}{2} K(x - x_0)^2 \quad (4.3)$$

$$dU = V_{\text{DC}} dq + F_{\text{ext}} dx \quad (4.4)$$

Here, V_{DC} is the potential difference between the electrode plates, F_{ext} is the external net force and q is the charge stored in the transducer. The Legendre transform U' can be used to calculate F_{ext} and q , in the present case of a voltage controlled electrostatic transducer [130]. U' and its partial differential are given by:

$$U' = U - qV_{\text{DC}} \quad (4.5)$$

$$dU' = F_{\text{ext}} dx - qdV_{\text{DC}} \quad (4.6)$$

F_{ext} and q can be calculated from the partial differential of the Legendre transform:

$$F_{\text{ext}} = \left(\frac{dU'}{dx} \right) = \frac{V_{\text{DC}}^2 \varepsilon_0 A_p}{2x_e^2} + K(x - x_0) \quad (4.7)$$

$$q = -\left(\frac{dU'}{dV_{\text{DC}}}\right) = \frac{\varepsilon_0 A_p V_{\text{DC}}}{x_e} \quad (4.8)$$

The total differential of F_{ext} and q are given by:

$$dF_{\text{ext}} = \left(-\frac{\varepsilon_0 A_p V_{\text{DC}}^2}{x_e^3} + K\right) dx + \frac{\varepsilon_0 A_p V_{\text{DC}}}{x_e^2} dV_{\text{DC}} \quad (4.9)$$

$$dq = -\frac{\varepsilon_0 A_p V_{\text{DC}}}{x_e^2} dx + \frac{\varepsilon_0 A_p}{x_e} dV_{\text{DC}} \quad (4.10)$$

For a given DC bias, the top electrode will move to the lowest-energy position, after which F_{ext} will be zero. If an additional AC voltage is applied to the transducer, the small signal may also influence the position of the top electrode. The small-signal AC voltage with angular frequency ω will induce two mechanical variables: a net force on the top electrode (F_{ext}) and a velocity of the top electrode (v); and two electrical variables: voltage on the top electrode (V) and current (i). Those four variables and their differential equations are expressed in eq. 4.11 and eq. 4.12. $\hat{F}$, $\hat{v}$, $\hat{V}$ and $\hat{i}$ are the complex amplitudes of the oscillating force, velocity, voltage and current respectively. α is the viscous damping factor.

$$F_{\text{ext}} = \hat{F}e^{j\omega t}; \quad v = \hat{v}e^{j\omega t}; \quad V = V_{\text{DC}} + \hat{V}e^{j\omega t}; \quad i = \hat{i}e^{j\omega t} \quad (4.11)$$

$$\frac{dF_{\text{ext}}}{dt} = j\omega\hat{F}e^{j\omega t}; \quad \frac{dx}{dt} = v = \hat{v}e^{j\omega t}; \quad \frac{dV}{dt} = j\omega\hat{V}e^{j\omega t}; \quad \frac{dq}{dt} = i = \hat{i}e^{j\omega t} \quad (4.12)$$

According to the second axiom of Newton,

$$m\frac{dv}{dt} = -F_{\text{ext}} - \alpha v \quad (4.13)$$

Inserting eq. 4.12 into dF_{ext} (eq. 4.9), dq (eq. 4.10) and Newton's axiom (eq. 4.13), we obtain the following equation system with four variables,

$$j\omega\hat{F} = \left(-\frac{\varepsilon_0 A_p V_{\text{DC}}^2}{x_e^3} + K\right)\hat{v} + \frac{\varepsilon_0 A_p V_{\text{DC}}}{x_e^2} j\omega\hat{V} \quad (4.14)$$

$$\hat{i} = -\frac{\varepsilon_0 A_p V_{\text{DC}}}{x_e^2} \hat{v} + \frac{\varepsilon_0 A_p}{x_e} j\omega\hat{V} \quad (4.15)$$

$$\hat{F} = -(j\omega m + \alpha)\hat{v} \quad (4.16)$$

By solving the equation system above (eq. 4.14, eq. 4.15 and eq. 4.16), the device impedance at up-state is found to follow [130]

$$Z_{\text{up}} = \frac{\hat{V}}{\hat{i}} = \frac{1}{j\omega C_p} \times \frac{\omega^2 m - j\omega\alpha + \frac{\varepsilon_0 A_p V_{\text{DC}}^2}{x_e^3} - K}{\omega^2 m - j\omega\alpha - K} \quad (4.17)$$

It is clear from eq. 4.17 that the DUT is far from purely parallel-plate capacitance, as usually assumed (and observed at radio-frequencies). The impedance shows a pole at $\omega^2 = \frac{K}{m}$. The impedances of the up-state switch at high frequency ($Z_{\text{up,HF}}$, $\omega^2 \gg \frac{K}{m}$) and low frequency ($Z_{\text{up,LF}}$, $\omega^2 \ll \frac{K}{m}$) following from eq. 4.17 are given by:

$$Z_{\text{up,HF}} = \frac{1}{j\omega C_p} \times \frac{\omega^2 m}{\omega^2 m} = \frac{1}{j\omega C_p} \quad (4.18)$$

$$Z_{\text{up,LF}} = \frac{1}{j\omega C_p} \times \frac{\frac{\varepsilon_0 A_p V_{\text{DC}}^2}{x_e^3} - K}{-K} = \frac{1}{j\omega C_p} \times \left(1 - \frac{\varepsilon_0 A_p V_{\text{DC}}^2}{x_e^3 K} \right) \quad (4.19)$$

Both limit cases are purely capacitive, but the capacitance at low frequency is larger than the parallel-plate capacitance, as indeed experimentally observed (cf. Fig. 3.12 and Fig. 3.15). Moreover, the capacitance difference far from resonance is frequency independent, as indeed observed. The DUT shows a significant mechanical resonance at ~ 17 kHz at zero DC bias, as confirmed with Laser Doppler vibrometer measurements. When the frequency approaches this frequency, the real part of the impedance increases (more details are given in section 4.1.2). The difference between eq. 4.18 and eq. 4.19 can be directly calculated because K can be estimated from V_{pi} [130],

$$K = \frac{27\varepsilon_0 A_p V_{\text{pi}}^2}{8\left(x_0 - x_d + \frac{x_d}{\varepsilon_r}\right)^3} \quad (4.20)$$

Replotting the measurements in Fig. 3.15 at LF and HF to isolate the additional term in 4.19 yields a quantitative match between the observed and predicted capacitances, as shown in Fig. 4.2 - without any parameter fitting. Since the capacitance difference far from resonance is frequency independent, the up-state LF capacitance in Fig. 4.2 can represent the capacitance at frequency below 10 Hz, and HF capacitance can represent the capacitance at frequency above 100 kHz. The C - V curves measured between 10 kHz and 100 kHz are discussed in the next section.

Figure 4.2: Measured and modeled ratio of LF capacitance (C_{LF}) to HF capacitance (C_{HF}) at up-state versus DC bias.

In short, the capacitance measured at both LF and HF is determined by the DC bias. The parallel-plate capacitance is measured at HF. However, the top electrode displaces with AC voltage at LF and contributes to the measured AC current, resulting in an increase of measured capacitance. Therefore, the measured LF capacitance is larger than the HF capacitance.

4.1.2 Around resonance frequency

The transducer model in section 4.1.1 explains the difference between the up-state LF and HF C - V curves above 100 kHz. We are curious about what happens to the up-state C - V curve when the measurement frequency approaches the resonance frequency.

Because of limitations of our measurement equipment, the open impedance ($> 10^6 \Omega$) is too large to be accurately measured below 100 kHz using the HF capacitance measurement technique; thus it is difficult to obtain an accurate up-state capacitance (~ 0.5 pF). Some HF C - V curves measured below 100 kHz are shown in Fig. 4.3. According to the analysis in chapter 3, the open compensation is DC bias independent. The open capacitance connects in parallel with C_{DUT} . The HF C - V curves below 100 kHz may shift in the y direction because of the inaccurate open capacitance, but the shape of the C - V curves do not change.

By comparing the difference between up-state capacitance at -20 V (C_{-20V}) and 0 V (C_{0V}), it is possible to see how the shape of up-state C - V curve changes when the frequency is approaching the resonance. From

Figure 4.3: The C - V curves measured below 100 kHz using the HF capacitance measurement technique; DC bias sweeps from 0 to -25 V.

Figure 4.4: The difference between up-state DUT capacitance at -20 V (C_{-20V}) and 0 V (C_{0V}) as a function of frequency.

Fig. 4.4, the capacitance difference ($C_{-20V} - C_{0V}$) firstly decreases then increases with the decrease of measurement frequency.

The tendency of the capacitance-frequency relation in Fig. 4.4 may be explained by the transducer model in 4.1.1. The capacitance equals to C_p at any frequency (except exactly at $\omega = \sqrt{\frac{K}{m}}$) when the DC bias is zero. At the other DC bias, the capacitance deviates from C_p around the

resonance frequency, which may show a maximum value and a minimum value according to eq. 4.17. Fig. 4.5 shows the impedance-frequency curve and capacitance-frequency curve calculated from eq. 4.17 at DC bias of 18.5 V. When the damping factor is small, the maximum impedance appears at frequency of f_1 in eq. 4.21 and the minimum impedance appears at frequency of f_2 in eq. 4.22. With the increase of the damping factor, the maximum impedance decreases whereas the minimum impedance increases, as shown in Fig. 4.5 (a). When the damping factor is large enough, the impedance peaks at resonance disappear (the system is also called an over-damped system).

$$f_1 = \frac{\sqrt{\frac{K}{m}}}{2\pi} \quad (4.21)$$

$$f_2 = \frac{\sqrt{\frac{K - \frac{\varepsilon_0 A_p V_{DC}^2}{x_e^3}}{m}}}{2\pi} \quad (4.22)$$

The capacitance is in ratio of the reciprocal of impedance, thus the capacitance firstly decreases then increases with a decreasing frequency, as shown in Fig. 4.5 (b). Therefore, the modeled capacitance-frequency curve has the same shape (tendency) as our measurement results.

To be noticed, although our 1-D model is able to predict the shape (tendency) of capacitance-voltage or capacitance-frequency curves around resonant frequency, it cannot accurately model the behavior around resonant frequency. Therefore, this simple 1-D model is only suitable to be used at frequency far away from the resonant frequency.

4.1.3 Artefacts at pull-in and pull-out

The above analysis demonstrates that the up-state capacitance measured around or below resonance frequency is influenced by the mechanical characteristics of DUT. In section 4.1.3, we show that the artefacts at V_{pi} and V_{po} in LF and QS C - V curves, as marked with oval circles in Fig. 3.15, are also because of the mechanical displacement of the top electrode.

The prominent spikes in the QS C - V curve occur at the two points of mechanical switching (pull-in and pull-out). Here, the capacitance abruptly changes. To maintain approximately the same voltage, an extra current must be supplied. This is well captured by a quasi-static measurement which measures current continuously, and is converted to capacitance. The following equations express how the QS technique measures a voltage dependent capacitance: the voltage change from V_1 to $V_1 + V_{step}$ and the

(a)

(b)

Figure 4.5: Impedance-frequency curve and capacitance-frequency curve calculated from eq. 4.17 around resonance frequency at DC bias of 18.5 V.

parallel-plate capacitance change from C_1 to $C_1 + \Delta C$, resulting in a change of charge from Q_1 to Q_2 .

$$Q_1 = C_1 V_1 \quad (4.23)$$

$$Q_2 = (C_1 + \Delta C) \times (V_1 + V_{\text{step}}) \quad (4.24)$$

According to the QS measurement principle in section 3.1.2.1, the measured capacitance is

$$C_m = \frac{Q_2 - Q_1}{V_{\text{step}}} = C_1 + \Delta C + \frac{\Delta C V_1}{V_{\text{step}}} \quad (4.25)$$

As in eq. 4.25, the measured capacitance is $\frac{\Delta C V_1}{V_{\text{step}}}$ larger than the real capacitance. At pull-in and pull-out, a large ΔC is amplified by $\frac{V_1}{V_{\text{step}}}$, resulting in spikes at pull-in and pull-out. This element of $\frac{\Delta C V_1}{V_{\text{step}}}$ can also explain why QS capacitance is larger than the parallel-plate capacitance when the DC bias is approaching V_{pi} .

The spikes at pull-in and pull-out in the LF C - V curve are sometimes present with varying significance, and are usually smaller than the spikes in the QS C - V curve. As these spikes are transient, and the LF curve is produced by sampling at intervals, measurement timing details determine the appearance of such spikes in the LF C - V curve.

4.2 Transducer model of the down-state capacitance

The model of down-state capacitance is more complicated than that of up-state capacitance because of the contact of top electrode and dielectric layer, making some main parameters in the model DC bias dependent.

4.2.1 Application of the transducer theory to the down-state capacitance and its limitation

If we describe the down-state with the same one-dimensional model as the up-state, we must account for the additional contact force exerted by the dielectric film. An elastic contact happens when the gap between the top electrode and the dielectric layer satisfy $2\sigma < x - x_d < 3\sigma$ [26]. As shown in Fig. 4.1, the height distribution of the under-surface of the top electrode is described by a Gaussian distribution, which has a standard deviation σ and a midplane (zero point) at $x = 3\sigma + x_d$ [26]. The height distribution is:

$$\phi(x) = \frac{1}{\sigma\sqrt{2\pi}} e^{-\frac{(x-x_d-3\sigma)^2}{2\sigma^2}} \quad (4.26)$$

Because of the surface roughness, the real contact area is [26]:

$$A_c = A_p \int_{3\sigma}^{\infty} \phi(x) dx = \frac{A_p}{2} \operatorname{erfc}\left(\frac{x - x_d}{\sqrt{\pi}\sigma}\right) \quad (4.27)$$

Using the indentation hardness H as an approximation of the elastic hardness, the contact force follows eq. 4.28 [26]. The hardness $H = 3.5 \times 10^8 \text{ N/m}^2$ is measured with a Berkovich tip in the previous study of DUT [26]. The contact force is given by

$$F_c = H A_c = \frac{H A_p}{2} \operatorname{erfc}\left(\frac{x - x_d}{\sqrt{\pi}\sigma}\right) \quad (4.28)$$

Thus the external net force (F_{ext}) applied on the top electrode is the sum of the spring force (F_s), the electrostatic force (F_e) and the contact force. The net force and its partial differential are given by:

$$F_{\text{ext}} = F_s + F_e + F_c = K(x - x_0) + \frac{\varepsilon_0 A_p V_{\text{DC}}^2}{2x_e^2} - \frac{H A_p}{2} \operatorname{erfc}\left(\frac{x - x_d}{\sqrt{\pi}\sigma}\right) \quad (4.29)$$

$$dF_{\text{ext}} = \left(K - \frac{\varepsilon_0 A_p V_{\text{DC}}^2}{x_e^3} + \frac{H A_p e^{-\left(\frac{x-x_d}{\sqrt{\pi}\sigma}\right)^2}}{\pi\sigma} \right) dx + \frac{\varepsilon_0 A_p V_{\text{DC}}}{x_e^2} dV_{\text{DC}} \quad (4.30)$$

If we define an effective spring constant at down-state K' as written in eq. 4.31 and replace the K in the dF_{ext} at up-state (eq. 4.9) by K' , we will get the dF_{ext} at down-state (eq. 4.30).

$$K' = K + \frac{H A_p e^{-\left(\frac{x-x_d}{\sqrt{\pi}\sigma}\right)^2}}{\pi\sigma} \quad (4.31)$$

According to study of H. M. R. Suy et al. on the same samples [26], the parallel-plate capacitance still can be approximated as shown in eq. 4.1 when $2\sigma < x - x_d < 3\sigma$. So the charge stored on the capacitance and its partial differential yield again eq. 4.8 and eq. 4.10.

Therefore, similar to the analysis in the up-state, the down-state DUT impedance, as written in eq. 4.32, is the same to the up-state DUT impedance with a higher spring constant K' to account for the mechanical contact and a different damping factor. To complicate matters, all model parameters become strongly gap and (therefore) voltage dependent.

$$Z_{\text{down}} = \frac{1}{j\omega C_p} \times \frac{\omega^2 m - j\omega\alpha + \frac{\varepsilon_0 A_p V_{\text{DC}}^2}{x_e^3} - K'}{\omega^2 m - j\omega\alpha - K'} \quad (4.32)$$

The down-state impedance shows a pole (resonance frequency) at $\omega^2 = \frac{K'}{m}$. The K' is much larger than K , which explains why the resonant frequency of DUT at down state is much larger than that at up-state. The associated down-state first resonance frequency is in the few-MHz range, as established by a 3D FEM calculation of DUT [131]. This explains the observed (small) difference between HF and RF capacitances in the down-state. The impedances of the down-state switch at radio frequency ($Z_{\text{up,RF}}$, $\omega^2 \gg \frac{K'}{m}$) and low frequency ($Z_{\text{up,LF}}$, $\omega^2 \ll \frac{K'}{m}$) follow from eq. 4.32 and are given by:

$$Z_{\text{down,RF}} = \frac{1}{j\omega C_p} \quad (4.33)$$

$$Z_{\text{down,LF}} = \frac{1}{j\omega C_p} \times \left(1 - \frac{\varepsilon_0 A_p V_{\text{DC}}^2}{x_e^3 K'} \right) \quad (4.34)$$

See Fig. 4.6, the fitted value of σ is 16.08 nm, in reasonable agreement with the 21.7 nm of standard deviation of the surface roughness measured by AFM on the top electrode. The damping factor is important to explain the HF C - V behavior, but does not influence the modeling of the LF and RF capacitance. The fitted and measured ratio of C_{LF} to C_{RF} have the same tendency versus DC bias, but they do not match each other well from -13 V to V_{po} . One possible reason is that the contact area A_c is not accurate. The top electrode may be curved, so the parallel-plate area A_p in eq. 4.27 should be replaced by the area of the contacted part of the plate which is a function of DC bias, $A(V_{\text{DC}})$. Since the down-state capacitance deviates from device to device when the DC bias is approaching V_{po} (rf. Fig. 3.14 (a)), the contacted part probably also deviates from device to device. Besides, adhesion force, which is not considered in the model, may also play a role and contribute to the difference between measurements and modeling results. Indeed, we can introduce more variables to $A(V_{\text{DC}})$ to make the data fitting better. However, each device has different fitted values of the variables, so it is not meaningful to do so.

The one-dimensional model leading to eq. 4.32 however cannot explain the observed differences between LF and HF in the down-state, because the HF range is below the first resonance frequency of the down-state switch.

Figure 4.6: Measured and fitted ratio of LF capacitance (C_{LF}) to HF capacitance (C_{HF}) at down-state versus DC bias.

4.2.2 Squeeze film effect on down-state capacitance

To explain the HF C - V curve, it is necessary to take into account that neither top electrode nor dielectric are perfectly flat. In the down-state, the movement of the top electrode is constrained by the points of contact to the dielectric on the bottom electrode. Away from those points of contact, air gaps remain under the top electrode - in particular near the springs. Parts of the top electrode may therefore still move upon an electrostatic force, contributing to the AC current, and change of measured capacitance. This is visible in a vibrometer, as shown in the inset of Fig. 4.7. Part of the top electrode (mainly the red part in the insert of Fig. 4.7) still vibrates with the AC signal in the down-state.

At atmospheric pressure, the top electrode movement is hampered by the squeeze film effect [132]. Mechanical motion of the top electrode compresses the partially enclosed gas volume between the electrodes. Given enough time, this pressure change leads to a lateral gas flow. As a consequence, this leads to viscous damping at very low frequencies, and an additional spring force (squeeze film stiffness) at higher frequencies. The FEM model of such devices [131] predicts a transition around 5 kHz for this effect.

Intuitively, one can argue that this effect should be most prominent when the applied field is low. Near pull-out, the top electrode is close to releasing. Therefore, the top electrode is relatively free to move and the

Figure 4.7: The difference between down-state DUT capacitance at -25 V (C_{-25V}) and -9.5 V ($C_{-9.5V}$) as a function of frequency. The inset shows the vibration amplitude of the top electrode in down-state measured at reduced pressure ($< 3 \times 10^{-4}$ mbar) at 1.5 MHz.

gas volume between the electrodes may be significant. With the decrease of frequency, the top electrode movement gradually appears while gas compression decreases, resulting in an increase of the measured capacitance near pull-out, as shown in Fig. 4.7.

4.2.3 Measurements in vacuum

The squeeze film effect is significantly reduced if the DUT is operated in vacuum. As a result, much higher vibration amplitudes are observed near pull-out (DC bias of 12 V) in vacuum, as shown in Fig. 4.8 (a). In the vibrometer measurement, we measure the vibration amplitude of an array of positions across the top electrode. The y-axis ‘displacement’ in Fig. 4.8 is the average vibration amplitude of all positions. When the DC bias is far from V_{po} , the top electrode hardly displaces in both atmospheric pressure and vacuum because of the limited space between the top electrode and the dielectric layer on bottom electrode. Thus, at DC bias of 25 V, the vibration amplitude is close to zero under both vacuum and atmospheric pressure. As shown in Fig. 4.8 (b), the vibration amplitudes in vacuum significantly increase when the DC bias is approaching pull-out, whereas the vibration amplitude under atmospheric pressure remains low at the down-state.

To further confirm the role of squeeze film damping, we carried out a HF C - V measurement in a vacuum chamber, see Fig. 4.9. The high-frequency,

(a)

(b)

Figure 4.8: Vibrometer measurement in atmospheric and low pressure: (a) displacement-frequency relation at DC bias of 12 V and 25 V; (b) displacement-DC bias relation at 1 MHz.

Figure 4.9: HF C - V curve at 100 kHz at atmospheric and low pressure. The setup used to this purpose did not allow for de-embedding, leading to a slightly higher overall capacitance compared to the C_{DUT} shown before.

low-pressure result shows a similar capacitance behavior in the down-state as the LF capacitance under atmospheric pressure.

In the down-state, each device will behave differently because of microscopic physical differences. This leads to a large device-to-device spread in the LF capacitance close to pull-out. This spread is expected to occur also for low-pressure measurements at higher frequency (>100 kHz). The down-state capacitance near pull-out is therefore not well controlled and this operating condition should be avoided.

4.3 Conclusion

A transducer model is introduced to explain the observed C - V curves described in chapter 3: the LF/QS capacitance (< 10 Hz) is up to 200% higher around the pull-in voltage and 25% higher around the pull-out voltage than the HF/RF capacitance (>100 kHz). In the up-state, this can be attributed to the mechanical movement of the top electrode at LF. A one-dimensional spring-mass model quantitatively reproduces the up-state behavior. This model is also able to explain the difference between RF and LF capacitances in the down-state with a fitted surface roughness. The difference between LF and HF capacitances in the down-state is attributed to squeeze-film effects in the residual gas volume between the top electrode and the dielectric film. This assumption is supported by vibrometer measurements, FEM simulations, and C - V measurements at reduced pressure. The LF capacitance measurement detects the coupling between mechanical

and electrical operation of RF MEMS switches, and helps to construct a comprehensive understanding of the device. A concise impedance model for RF MEMS capacitive switches should include these properties, in particular for biasing voltages close to pull-in and pull-out.

Chapter 5

Spring constant measurements of RF-MEMS switches

Some important parameters of the RF-MEMS capacitive switches, such as the pull-in voltage (V_{pi}) and the resonance frequency, relate with the spring constant (k). The measurement of k is important to characterize the RF-MEMS switch; it may be a suitable parameter for in-line process control if a fast and reliable wafer-level electrical measurement is developed.

In this chapter, we discuss various approaches to determine the spring constant in RF MEMS capacitive switches. Three approaches are tested on four designs of RF-MEMS switches, and compared with 3D FEM simulation results of the same devices. In section 5.1, we use the lowest vibration mode, as obtained from vibrometry, the pull-in voltage (V_{pi}), and the low-field capacitance-voltage ($C-V$) curve of the device to extract the spring constant. Other possible methods to obtain the spring constant are also introduced. In section 5.2, experimental results are presented for each approach, and FEM model predictions are used to further verify and interpret the findings.

5.1 Theories of various methods for spring constant measurements

In this section, we describe various approaches to derive the spring constant from device measurements; advantages and disadvantages of each method are discussed.

5.1.1 Spring constant from the vibrometry measurement

A measurement of the mechanical resonance behavior of the top electrode under periodic AC bias can be used to determine the spring constant. At low frequencies, the top electrode displaces with the AC bias, and this displacement has a phase difference with the input of the AC bias. Both amplitude and phase of the displacement depend on the applied frequency. The phase difference between the electrical input and the mechanical displacement output equals -90° at the eigen frequency ω_0 associated with the resonance. The resonance frequency is the frequency at which the maximum displacement of the top electrode occurs. The resonance frequency is close to ω_0 but becomes smaller than ω_0 for damped devices [130]. Therefore, we use the eigen frequency to calculate k . Vibrometry is used to measure this resonance behavior. In the (simplified) one-dimensional approximation of a mass-spring system we have:

$$k = \omega_0^2 m = \omega_0^2 \rho A_p t_{TE} \quad (5.1)$$

In this equation, m is the effective mass of the top electrode which is approximated with the mass of the rectangular electrode plate, ρ is its mass density, A_p is its effective surface area, and t_{TE} is its thickness [5]. We can see from this equation that the accuracy of the mass estimation affects the absolute value of k linearly.

This vibrometry method has the advantage of being independent of the gap size, and insensitive to fringe capacitance; plus it does not require de-embedding. The dedicated measurement set-up needed for this test can only be used for unpackaged devices.

5.1.2 Spring constant from the pull-in voltage

Using the dimensions in Fig. 4.1, we define the effective gap of the parallel-plate capacitance at zero DC bias: $g = x_0 - x_d + \frac{x_d}{\epsilon_r}$, and the vertical displacement of the top electrode from the equilibrium position at zero DC bias: $u = x - x_0$. Thus $g + u = x_e = x - x_d + \frac{x_d}{\epsilon_r}$ is the bias dependent effective gap of the parallel-plate capacitance. As described in chapter 4, assuming that the electrostatic force can be described through a parallel-plate approximation, and that Hooke's law is applicable, we can find the following equation for the equilibrium of the top electrode under any DC bias:

$$F_{\text{ext}} = \frac{V_{\text{DC}}^2 \epsilon_0 A_p}{2(g + u)^2} + ku = 0 \quad (5.2)$$

When the vertical displacement of the top electrode goes up to one third of the effective gap, namely $g + u = \frac{2}{3}g$, pull-in occurs [130]. Inserting $V_{\text{DC}} = V_{\text{pi}}$ and $g + u = \frac{2}{3}g$ into eq. 5.2, the spring constant is found as

$$k = \frac{27\varepsilon_0 A_{\text{p}} V_{\text{pi}}^2}{8g^3} \quad (5.3)$$

With the measurement of V_{pi} (for instance, with a high-frequency C - V measurement), and knowledge of the geometrical parameters, we can thus determine k . As pull-in has a very pronounced signature in the C - V curve, V_{pi} determination is easy and accurate, even without calibration and de-embedding. Measurement-induced stress effects may occur using this method because of the applied high DC bias, in contrast to the other two approaches.

In the above equations we assume that no dielectric charge is present. A homogeneous dielectric charge would lead to a voltage shift V_{s} of the entire C - V curve along x -axis. So one may take the average of (the absolute value of) the positive and negative V_{pi} to avoid the influence of dielectric charge. We have further assumed that the fringe capacitance is negligible, i.e. the parallel-plate equation can be used for the capacitance in eq. 5.2. If the fringe capacitance is considered, we can, as first order correction, treat this as an area correction in eq. 5.2 and eq. 5.3. As a result, the determination of k is linearly affected by fringe capacitance errors. Finally, as the top electrode is not rigid and has several degrees of motional freedom, such as inducing intermediate state, which may yield inaccurate results in this approach.

5.1.3 Spring constant from the low-field C - V curve

We can also estimate the spring constant from the “sag” of the top electrode when a small DC bias is applied, by measuring its capacitance around 0 V, with a measurement frequency much larger than the eigen frequency. At low-field, $u \ll g$, and therefore the parallel-plate capacitance can be approximated as

$$C_{\text{p}} = \frac{\varepsilon_0 A_{\text{p}}}{g + u} \simeq \frac{\varepsilon_0 A_{\text{p}}}{g} \left(1 - \frac{u}{g}\right) \quad (5.4)$$

From eq. 5.2 follows that

$$u = -\frac{V_{\text{DC}}^2 \varepsilon_0 A_{\text{p}}}{2k(g + u)^2} \simeq -\frac{V_{\text{DC}}^2 \varepsilon_0 A_{\text{p}}}{2kg^2} \quad (5.5)$$

Inserting eq. 5.5 into eq. 5.4:

$$C = \frac{\varepsilon_0^2 A_p^2}{2kg^4} V_{DC}^2 + \frac{\varepsilon_0 A_p}{g} \quad (5.6)$$

So the low-field C - V curve can be approximated by a parabola shape: $C = aV_{DC}^2 + b$ as is experimentally observed earlier [82]. The minimum capacitance equals b . The spring constant can be determined as

$$k = \frac{\varepsilon_0^2 A_p^2}{2g^4 a} \quad (5.7)$$

In eq. 5.6, V_{DC} should be replaced by $V_{DC} - V_s$ if homogeneous dielectric charging is considered, while eq. 5.7 remains unaffected. Clearly, when we use this equation we are more sensitive to dimensional errors, given the quadratic dependence on A_p and even a fourth power dependence on g .

5.1.4 Other methods to obtain spring constant

Since our devices can be simplified to a 1-D mass-spring model (as described in chapter 4), the spring constant can also be calculated directly from the applied mechanical force divided by the vertical displacement of the movable top electrode. We were not able to do experiments of this method because of the lack of an appropriate measurement equipment in our labs to directly apply mechanical force.

It is possible to find an alternative method to apply force and calculate spring constant. Beside applying mechanical force, we can apply voltage and provide electrostatic force between the two electrodes. The value of electrostatic force can be well controlled by the amplitude of applied voltage.

There are three ways to measure the displacement of the movable top electrode: mechanical, optical and electrical. The mechanical way is to contact a cantilever with a tip needle to the movable electrode and measure its displacement. Our labs do not have the appropriate equipment to do this measurement. The optical way and electrical way both indirectly measure the displacement: the former one detects displacement from the change of light reflection, and the later one calculates displacement from the change of the parallel-plate capacitance. We did not choose these methods because of the poor measurement accuracy.

Besides the difficulties in experiments as mentioned above, Hooke's law is related to linear deformation. It is possible that the deflection is too large to ensure a linear displacement-force relation. Therefore, it is not suitable to calculate the spring constant directly from Hooke's law, no matter how we apply force and how we measure the displacement.

Figure 5.1: Top view microscope image of the RF-MEMS capacitive switches under study.

5.2 Result and discussion on spring constant measurements

The measurement results of each approach are discussed and compared in this section.

5.2.1 Experiments

A top view microscope image of the four studied RF-MEMS capacitive switches is shown in Fig. 5.1. The devices are supplied by EPCOS AG; they are produced in an intermediate stage of process development and intended for reliability characterization. Further documentation of the switches is found in [48]. The switches have various spring designs, with longer springs going from device 0 to device 3. Except for the length of the springs, the dimensions and materials are identical in the four switches. More details of the devices can be found in chapter 3 and chapter 4, since device 3 is the device under test in those two chapters.

The mechanical behavior is measured by scanning laser Doppler vibrometry. A DC bias and a sinusoidal AC voltage is applied to the MEMS switch; the velocity of the top electrode and the phase between the electrical input and the mechanical output are measured. The mechanical displacement of the top electrode is calculated by velocity integration.

The C - V curve is measured with a Keithley 4200 CVU at 800 kHz. The

Figure 5.2: Displacement-frequency relation of device 3 at (a) AC voltage of 0.1 V and various DC biases; and (b) at DC bias of 9 V and various AC voltages.

open de-embedding is done as described in chapter 3. The short compensation is negligible at 800 kHz. In addition, each MEMS switch under study is switched twice before characterization, as explained later in section 5.2.2.

The measurement results of the MEMS switches will be compared with a 3D multi-physics finite-element simulation performed by Jeroen Bielen of EPCOS NL, which has been set up to model the capacitance-voltage (C - V) behavior of the switches in ANSYS and validated on data from similar devices as studied in this work [131, 133].

5.2.2 Vibrometry

Fig. 5.2 shows the mechanical response of the RF MEMS switches as measured by a vibrometer. As shown in Fig. 5.2 (a) and (b), the magnitude of displacement of the top electrode increases with DC bias and with AC voltage, consistent with a larger u leading to a higher electrostatic force. The resonance frequency (and eigen frequency) is independent of both DC bias and AC voltage within the measured range. In principle, the MEMS switch is considered a voltage controlled electrostatic transducer. Its eigen frequency will reduce when the DC bias approaches the V_{pi} , reflected in a lower effective spring constant [130]. However, when the DC bias is not close to V_{pi} , the decrease of eigen frequency is not significant, and the estimation of spring constant in eq. 5.1 is still accurate.

To obtain low-noise measurement curves, we use a DC bias of 9 V and an AC voltage of 1 V to study the four MEMS switches. The displacement-frequency and phase-frequency curves of the four MEMS switches under

Figure 5.3: The (a) displacement and (b) phase as a function of frequency of four MEMS switches, measured at DC bias of 9 V and AC voltage of 1 V.

study are shown in Fig. 5.3 (a) and (b). The resonance frequency (where the maximum displacement happens) is slightly smaller than the eigen frequency (where the phase equals to -90°). It is significantly lower when the springs are longer, as expected.

The resonances shown in Fig. 5.2 and Fig. 5.3 are the first (fundamental) resonance. If we increase the frequency of the AC signal, further resonances will be measured, because the MEMS switch has more than one resonance mode [5, 131]. The spring constant given in eq. 5.1 is actually the modal spring constant of the fundamental mode.

5.2.3 Pull-in voltage

To investigate the extraction of k from V_{pi} , C - V curves are measured on the four devices. Table 5.1 shows the extracted V_{pi} of each device, as measured and according to the FEM model. The model reproduces the initial deformation effects and the resulting reduction in the V_{pi} . Model and measurement results reflect the stable device behavior after the first two switching cycles. Comparing the results in Table 5.1, the measured and modeled V_{pi} differ by 1.5% to 7%. In other words, the FEM model matches the measurement quite well. We use the V_{pi} (of both FEM model and measurement) to calculate the spring constant, as further discussed in section 5.2.4.

Table 5.1: Comparison of modeled and measured pull-in voltage.

V_{pi}	device 3	device 2	device 1	device 0
FEM model	20 V	44 V	49 V	52 V
Measurement	19.4 V	44.7 V	52.5 V	54.7 V

Figure 5.4: Five consecutive measurements for the open compensation of device 3, indicating the poor short term repeatability.

5.2.4 Low-fields capacitance method

From the same C - V measurements as in section 5.2.3, we can obtain k also from the low-field capacitance. The low capacitance in the up-state is a complicating factor here, because the parabola method relies on the precise determination of small capacitance changes. In the devices under test, both the minimum capacitance in the C - V curve and the open compensation are smaller than 1 pF and are in the lowest measurement range of the Keithley 4200 CVU, where the accuracy of the instrument is not specified in the current user manual.

Indicative of the situation is the short term repeatability of the open compensation measurement. Fig. 5.4 shows five consecutive measurements of an open structure, without any changes to the setup in between. Although each voltage sweep produces data with a very small rms value below 1% of the measured capacitance, which could be misinterpreted as an indication of a very accurate measurement, the consecutive sweeps yield results separated as much as 5% (sweeps 3 and 5).

Figure 5.5: Relation between $\frac{C_p}{C_{\min}}$ and DC bias, according to the FEM model. C_p is the voltage dependent parallel-plate capacitance; $C_{\min}$ is the minimum capacitance in the parabola.

This presumed drift is 0.031 pF which amounts to as much as 15% of the minimum capacitance of the MEMS switches under study. However, one may argue that the determination of the parabolic parameter a is rather insensitive to drift; so not all errors of the instrument will automatically affect the determination of k .

A second consideration in the determination of the spring constant from a is the chosen voltage range of the parabola measurement. If the range of DC bias is too large, the assumption that $u \ll g$ is imprecise. If the range of DC bias is too small, the C - V curve is noisy making the parabola fitting inaccurate. Fig. 5.5 shows the relation between $\frac{C_p}{C_{\min}}$ and DC bias according to the FEM model. $C_{\min}$ is the minimum capacitance in the C - V curve; C_p is the voltage dependent parallel-plate capacitance. In the FEM model calculations for this chapter, dielectric charging is set to zero, so $C_{\min}$ in the FEM data is the capacitance at zero DC bias. C_p increases slowly with DC bias when the DC bias is small; C_p increases dramatically when the DC bias approaches the V_{pi} .

Fig. 5.6 shows the outcome of the k -value obtained by the parabola method, for the 4 devices, and as calculated from FEM C - V data as well as measured C - V data. On the horizontal axis we plot the maximum value of the DC bias still included in the fit (so the fit is obtained on data from $-V_{\text{DC}}$ to $+V_{\text{DC}}$ in this figure). At lower DC bias ranges, the spring constant is correctly extracted, but at higher bias, the values are systematically off because of the deviation of C - V from the parabolic approximation (eq. 5.6).

Figure 5.6: Relation between normalized spring constant from low-field C - V parabola and the range of DC bias of the parabola, symbols for measurement and lines for FEM model; k_3 is the spring constant of device 3 calculated from the C - V parabola.

When the DC bias range is chosen too low, erroneous values are obtained on the experimental data, as a result of measurement and calibration errors.

The $\frac{k}{k_3}$ from the measured low-field C - V curve is larger than the $\frac{k}{k_3}$ as obtained from modeled low-field C - V data. The reason is that in our analysis, we assumed the four devices under study to have the same air gap (as being determined by the thickness of the sacrificial layer in the fabrication process). However, the springs of fresh devices may have plastic deformation after the first several switching cycles; so the air gap may change. This is particularly significant for the devices with short springs studied here, which are in fact designed for the purpose to quantify effects like this. The g in eq. 5.7 may increase because of the plastic deformation. Device 3 has the longest springs, which almost have no plastic deformation. The shorter the springs are, the more significant the plastic deformation is. These effects can be quantitatively reproduced by the FEM model in ANSYS, which is built by our project partner, as written in [131, 133]. A more precise k determination than presented here can be achieved by establishing g separately for each device.

In the further analysis we extract k values with this method using the following DC biases: 5 V, 15 V, 20 V and 20 V for device 3, device 2, device 1 and device 0 respectively.

Figure 5.7: Spring constants as obtained from three methods, normalized to the spring constant obtained on device 3 (k_3).

5.2.5 Comparison of the three methods

The spring constants calculated from the above three methods are shown in Fig. 5.7 and compared to the FEM result. All the obtained k are normalized to the spring constant of device 3 (k_3). All approaches yield a gradual increase of the spring constant when the springs are shorter, as expected; and quantitatively, a reasonable agreement is obtained. Only the k as obtained from the vibrometer deviates from the others. One possible reason is that the modal masses of the four devices are not accurately estimated from the dimensions of the plate.

We find that the parabola method is very sensitive to the deviation of g and A_p , and depends on small differences between small capacitance values. Although the normalized k matches the FEM model well, we can hardly get an accurate absolute value of k from the parabola method.

5.3 Conclusion

The spring constant of four RF-MEMS capacitive switch designs are extracted using vibrometry, pull-in voltage, and low-field C - V measurements. The obtained results are compared to a FEM model of the devices. The advantages and disadvantages of each method are described. The vibrometer results deviate from the other estimates. The low-field C - V method is able to easily and quickly detect the change of spring constant and the

relation of springs of various designs. However, the absolute value of spring constant obtained from the low-field C - V method is relatively inaccurate due to measurement noise and parameter uncertainties, which makes this method unsuitable for in-line process control. The spring constant from pull-in voltage is easy to measure and is accurate. To be noticed, the measurement frequency should be much higher than the eigen frequency. The disadvantage is that applying the high pull-in voltage to DUT may induce charging. This may not be a serious problem if the high voltage is applied once for the in-line process control. Therefore, we think the pull-in voltage method might suitable for in-line process control.

Chapter 6

Process influence on properties and reliability of PZT thin films

Galvanic (or ohmic/contact) RF-MEMS switches, especially the piezoelectrically actuated ones, show quite a few advantages over capacitive RF-MEMS switches as reviewed in chapter 2, hence our project aims to develop reliable piezoelectrically actuated galvanic RF-MEMS switches. The fabrication and reliability of piezoelectric materials is the key factor of the application of piezoelectric RF-MEMS switches as reviewed in chapter 2. $\text{PbZr}_{1-x}\text{Ti}_x\text{O}_3$ (PZT) is one of the most widely used piezoelectric materials in MEMS because of its excellent piezoelectric properties [134]. It has been reported that PZT properties are influenced by fabrication processing, especially plasma etching [135–144]. Nowadays, with the decreasing of device feature size, plasma etching (or dry etching) is more widely used than wet etching, because of its an-isotropic and accurate pattern transfer for PZT [145]. The further development of RF-MEMS switches application requires an increased integration density. Notwithstanding the benefits of integration, plasma etching induced dielectric damage may become a more prominent issue with increasing integration density [146–151]. Therefore, it is important to study the effects and mechanisms that cause plasma etching induced damage in PZT based devices, as a first step of the further development of our piezoelectrically actuated galvanic RF-MEMS switches.

In Chapter 6, two kinds of metal-PZT-metal capacitors are compared to study processing sequence induced changes on PZT. Section 6.1 introduces the fabrication of the two kinds of PZT capacitors. Section 6.2 compares the low-field current-voltage (I - V) curves of both kinds of virgin capacitors,

and discusses the process induces PZT self-poling. Section 6.3 reviews etching induced PZT damage, and gives a possible explanation of the difference between the two kinds of PZT capacitors in PZT permittivity and reliability.

6.1 Fabrication of two kinds of PZT capacitors under study

To study the plasma etching induced PZT self-poling phenomenon in section 6.2 and the process-induced PZT damage in section 6.3, we simultaneously fabricate two kinds of metal-insulator-metal (MIM) PZT capacitors on the same wafer, using the same materials, but make some changes to the the plasma etching process.

6.1.1 Fabrication processes

In the present work, the two kinds of MIM PZT capacitors are fabricated on TiO_2 terminated (001) oriented SrTiO_3 (STO) substrates. The schematic cross-section of the two kinds of PZT capacitors is shown in Fig. 6.1. Both kinds of capacitors have 80-nm-thick SrRuO_3 (SRO) bottom electrodes, and 800-nm-thick epitaxial $\text{PbZr}_{0.37}\text{Ti}_{0.63}\text{O}_3$ (PZT) thin films with (001) orientation. The PZT layer is deposited by pulsed laser deposition using a KrF Excimer laser (Lambda Physik, 248 nm wavelength) at a laser fluency of 2.5 J/cm^2 , substrate temperature of 600°C , pure oxygen pressure of 0.1 mbar, target-substrate distance of 6 cm, and 10 Hz repetition rate. More details of this PZT deposition are found in [152]. The SRO electrode is deposited as in [153].

The Au top electrode layer is deposited by sputtering. The square top electrode is defined by photolithography and then etched by Ar ion beam milling. A neutralizer is used during ion milling; so extra electrons are generated around the samples to compensate the positive charge of the Ar ions. In practice, the charge compensation usually is not perfect, thus this ion milling process may result in net charge on the etched samples. Firstly, a 20 nm Au layer is deposited and patterned by ion milling into four $4\times 4 \text{ mm}^2$ areas on the sample. The PZT outside the $4\times 4 \text{ mm}^2$ areas is also slightly exposed to the ion beam. After acetone clean, another 80 nm Au layer is deposited and patterned by ion milling into an array of $100\times 100 \text{ }\mu\text{m}^2$ squares across the sample. To be noticed: the latter ion milling uses a larger neutralizer current and a smaller ion beam current compared to the former ion milling, which may lead to higher possibility of

Figure 6.1: Schematic cross-section of two kinds of PZT capacitors in main process steps.

negative net charge on the etched samples. During the ion milling of this patterning step, the 20 nm gold film is also removed (unless where part of the $100 \times 100 \text{ } \mu\text{m}^2$ squares). The photoresist is also removed by acetone. This results in two kinds of capacitor structures: 1) the regular capacitors situated inside the $4 \times 4 \text{ mm}^2$ areas with a top electrode of $20 + 80 \text{ nm Au}$; and 2) the overetched capacitors outside the $4 \times 4 \text{ mm}^2$ areas with a top electrode of 80 nm Au . Due to the simultaneous ion-milling processing the latter devices are overetched into the PZT, while for the former devices the PZT is only lightly exposed to the ion beam, as shown in the schematic cross-section in Fig. 6.1. The top electrode-PZT interface of the overetched capacitor is rougher and more defective than that in the regular capacitor, because of the first ion milling process [144]. These two kinds of MIM PZT capacitors are made on the same wafer, have the same geometry and consist of the same materials.

6.1.2 Plasma etching principle

In order to study the plasma etching induced PZT damage/degradation, it is necessary to understand the working principle of plasma etching. Some plasma etching techniques such as RIE or ICP involve chemical reactions. For our experiments, a pure physical plasma etching method called Ar ion milling is used. The Ar ions contained in the plasma are accelerated by the ion source, and bombard the samples to be etched. The continuous bombardments result in the transfer of the kinetic energy to the thermal energy and heat the samples. The top layer atoms, atom clusters, or molecules of the samples can be driven away from the surface [154]. The main elements of the Ar ion milling technique are introduced in the following paragraphs.

A plasma reactor generates a plasma environment for the etching process. Inductively coupled plasma, capacitively coupled plasma and downstream plasma are three main plasma reactor types [155]. Usually, an inductively coupled plasma source generates a higher density of plasma than a capacitively coupled plasma source. The former plasma source is mostly used for conductor etching; and the latter plasma source is mostly used for dielectrics etching. A downstream plasma source is mostly used for photoresist stripping, dry cleaning or planar surface treatments [155].

Wide beam ion sources include gridded DC ion sources, gridded RF ion sources and RF sources with a single grounded grid [154]. The DC ion sources are used for our experiments. Electrons are generated from a filament by thermionic emission; Ar gas surrounds the filament; the electrons in turn ionize some of the Ar atoms and form a plasma. The Ar ions are attracted by a screen grid, and are accelerated by an accelerator grid

towards the samples to be etched. Because the ions repel each other, the ion beam may diverge. The divergence angle increases with the decrease of ion beam energy. To minimize ion divergence and sample charging, a neutralizing filament is used to generate electrons to neutralize the ions. The RF ion sources use inductively coupled RF power to ionize gas, usually for reactive gases such as O_2 . Both physical and chemical etching will take place if reactive gases are used. The RF sources with a single grounded grid use both capacitively and inductively coupled power to ionize gases. They generate quasi-neutral particles for etching without additional neutralizer, especially for low energy delicate etching [154].

A vacuum chamber is needed for the ion milling process. The vacuum level should ensure that the mean free path is longer than the distance between the ion sources and the samples to be etched. The mean free path is the average distance in which an ion, atom or molecule typically travels in vacuum before colliding with another particle. The direction of the ion, atom or molecule could be changed if it collides with another particle [154].

An end point detector is installed near the samples to be etched. The secondary ions coming from the sample surface are analyzed to determine the surface material and detect the end point of etching [154].

A sample holder usually has a water-cooling to prevent thermal damage to the samples because the ion bombardments continuously heat the samples. Some holders also have an extra heater to accelerate etching. Substrate heating gives energy to the surface particles, and allows them to be removed with less incident kinetic energy, resulting in an acceleration of the etching rate [154].

6.2 Process induced PZT self-poling

6.2.1 Polarization of PZT

To understand PZT poling and self-poling, the origin of PZT polarization is briefly introduced here. The Ti/Zr ion in a PZT unit cell has various stable positions. The position of the Ti/Zr ion determines the polarization of each PZT unit cell. The PZT unit cells can be considered to be electric dipoles. If a large external electric field is applied, all the Ti/Zr ions prefer to move to one stable position as shown in Fig. 6.2, in other words, the dipoles are aligned in one direction. Therefore, the PZT material shows polarization in the direction corresponding to the external electric field. The operation to align the PZT dipoles in one direction is called PZT poling.

Usually, the Ti/Zr ions are randomly located in the stable positions during PZT depositions, in other words, the Ti/Zr ions have equal chance to be

Figure 6.2: The unit cell of PZT material at two different external electric fields.

at each position. Therefore, the dipoles are considered to have random directions, and the average polarization of PZT can be considered zero, which means that the PZT material is not poled. However, in some situations, the Ti/Zr ions prefer to be in one stable position during processing, which is called process induced PZT self-poling [156]. The following subsections introduce the detection of PZT self-poling, the possible mechanisms of PZT self-poling, and give a possible explanation of the detected PZT self-poling in this study.

6.2.2 Switching current in low-field I - V curves

The process induced PZT self-poling can be detected in the low-field I - V measurements of virgin PZT MIM capacitors. Section 5.2.2 presents how the I - V measurements are conducted and how we detect the PZT poling direction from such measurements.

6.2.2.1 Low-field current measurements

All the electrical measurements in this chapter are conducted at 150 °C in dry nitrogen ambient by using a Keithley 4200 SCS. The bottom electrode and chuck are grounded in all the measurements. For low-field I - V measurements, the DC bias is applied to the top electrode using sweeps mode from zero to +8 or -8 V (as indicated by the arrows in Fig. 6.3) with a voltage step of 0.5 V. The timing of the current measurement is

Figure 6.3: I - V curves of two kinds of virgin PZT capacitors; arrows in the figure indicate the directions of the DC bias sweep.

critical for detection of the PZT switching current as described later in this section. The 4200 instrument measures a stable current during a measurement period determined by the sweep mode. In the presented study, we chose “quiet mode” operation to obtain an accurate current reading in the 100-pA regime. Two other factors also determine the overall timing of this measurement [157]. The instrument will observe a certain (unpredictable) settling time when bias conditions change. The settling time can be reconstructed from the overall sample-to-sample time interval. After this built-in settling time, a user-defined additional waiting time (termed sweep delay and set to 0 in this study) can be added between instrument settling and the measurement period. The time interval between consecutive data points in the I - V curve in this study is 0.8 s.

The I - V curves of both kinds of virgin PZT capacitors are shown in Fig. 6.3. The current shown in the figure is the absolute value of the measured current. For the regular PZT capacitor (with an $80 + 20$ nm Au top electrode), a current spike is detected at negative voltage but not at positive voltage. For the overetched PZT capacitor (with an 80 nm Au top electrode), it is the opposite: a current spike appears only during positive voltage sweep. In all cases, the switching current will not appear if we repeat the I - V measurement in the same direction.

6.2.2.2 Switching current and PZT self-poling

Switching current accompanies the large-scale switching of ferroelectric domains, which appears around the coercive fields [158]. As mentioned before, the time interval between consecutive data points in the I - V curve is an important parameter to detect the PZT switching current. People choose various time intervals for various purposes.

Usually, the I - V curve is used to study PZT leakage current. People usually set a small voltage step and a long delay time in I - V measurements to avoid the PZT switching current. For example, the time interval in I - V curves is at least 5 minutes in [159]; M. Nguyen uses voltage step of 0.1 V and delay time of 20 s for all his I - V measurements of the PZT material similar to this study. Therefore the current spike is absent in these I - V curves.

In this study, we choose a short time interval to detect the current spike and analyze the PZT poling. As described in [158], the low-field I - V measurement is an easy way to see whether the PZT material is poled at the polarity of the applied voltage sweep. If the PZT is positively, respectively negatively poled, the switching current can be detected at negative, respectively positive coercive fields. If the PZT layer is not poled, the switching current can be detected around both the positive and negative coercive fields.

The PZT switching time can vary from nanoseconds to hundreds of seconds [160]. Therefore, it is possible that our I - V measurements (~ 0.8 s of interval time between neighboring data points in I - V curves) detect part of the PZT switching current besides the leakage current of the PZT layer. If the sweep delay is increased, less or no PZT switching current should be detected. Indeed, when the sweeping delay is set to 10 s and the other parameters are kept the same, no current spikes are detected in the I - V curves of virgin devices, as shown in Fig. 6.4.

According to the first I - V measurements of virgin PZT capacitors as indicated by symbols in Fig. 6.3, for regular capacitors, the current spike appears at negative voltage but not at positive voltage; for overetched capacitors, the current spike appears at positive voltage but not at negative voltage. Therefore, the regular PZT capacitors are positively poled and the overetched PZT capacitors are negatively poled. When the I - V measurement is repeated in the same direction, the current spike should not appear because the PZT is poled in this direction during the previous I - V measurement, as indicated by solid lines in Fig. 6.3.

To understand why the regular and overetched PZT capacitors are poled in opposite direction during processing, we analyze the possible mechanisms

Figure 6.4: Comparison of low-field I - V curves measured in two different sweep delay conditions.

that cause PZT poling during processing in section 6.2.3 and 6.2.4.

6.2.3 Dislocations and PZT self-bias

6.2.3.1 The theory of PZT self-bias

Self-poling in PZT MIM capacitors has been observed by many authors [156, 161, 162]. Several research groups have proposed mechanisms to explain the self-poling in these stacks, as further treated in this section.

The self-bias of a MIM PZT capacitor is a well-studied reason to explain the self-poling of virgin capacitors. It means that the virgin MIM PZT capacitor behaves as if a voltage bias is applied to the capacitor. The self-bias is mainly determined by dislocations between the bottom electrode and the PZT layer and the dislocation related charges [156, 161].

The strain gradient could be an intrinsic mechanism of self-bias [161], as illustrated in Fig. 6.5 (a). Because of the lattice mismatch between bottom electrode and PZT, a strained layer is formed in PZT next to the bottom electrode. The strain is relaxed by dislocations, resulting in a strained PZT layer near bottom electrode and a non-strained PZT layer above. The strain gradient makes the strained PZT layer a non-switchable polar layer. M. Boota *et al.* provides a model to explain the formation of the non-switchable layer [161]. The polarization direction of the non-switchable layer is determined by the lattice difference between bottom electrode and PZT. The non-strained PZT layer is switchable, and prefers to follow the

Figure 6.5: Schematic build-up of PZT self-bias: (a) intrinsic mechanism, (b) extrinsic mechanism. The arrows indicate the dipoles in PZT.

direction of polarization in the non-switchable PZT layer [156]. The strain gradient also influences the grain size of PZT. In case of PZT on an STO substrate, the PZT has smaller grains near the bottom electrode-PZT interface compared to the upper PZT grain size.

The charged defects could be an extrinsic mechanism of self-bias [161], as illustrated in Fig. 6.5 (b). The strain relaxation in PZT accompanies the generation of charged defects. If the amount of defects is large enough, they can also influence the direction of polarization in the switchable PZT layer [156]. M. Boota *et al.* show that the intrinsic mechanism is stronger than the extrinsic mechanism [161].

The proposed mechanism described above suggests that the self-poling is caused by the lattice mismatch at the bottom electrode-PZT interface. It is therefore not to be expected that different top electrodes would lead to a difference in self-poling.

6.2.3.2 Influence of PZT self-bias in this study

In our study, the two kinds of PZT capacitors have difference interfaces between top electrode and PZT, but the top electrode-PZT interface hardly influences the PZT self-poling according to the theory of Boota et al.. Since these two kinds of PZT capacitors have exactly the same bottom electrode-PZT interface, we expect the same lattice mismatch effect, resulting in PZT self-poling in the same polarity. Therefore, PZT self-bias cannot explain the observed PZT self-poling in our study, and we look for some other reasons.

6.2.4 Plasma charging and PZT poling

6.2.4.1 Theory of plasma etching induced charging

Another possible reason of PZT self-poling is plasma induced charging. Common origins of plasma induced charging are the electron shading effect and the spatial non-uniformity of the plasma potential [150, 163].

The electron shading effect is induced by the different incident angular distributions between ions and electrons [150, 163]. The Debye sheath is the transition from a plasma surface to a solid surface. The ions are accelerated in the thin sheath thus have a narrow ion angular. Most of the ions bombard the wafer surface perpendicularly [164]. The electrons are decelerated in the sheath, resulting a broad electron angular distribution [164]. A schematic illustration of the electron shading effect is shown in Fig. 6.6 (a). When the pattern being etched has a large aspect ratio, the electrons may charge the photoresist. The negatively charged photoresist repels the light electrons. Thus the positive ions more easily reach the electrode being etched than the electrons, resulting in a positively charged electrode [147, 150]. Usually, the electron shading effect happens when the trench aspect ratio is larger than 2.4 [147], thus the trench aspect ratio of the studied PZT capacitors (<0.02) is small enough to avoid an electron shading effect. The extended electron shading effect happen during overetching of patterns with a low aspect ratio, as shown in Fig. 6.6 (b). Because most of the electrons have an oblique incident angle, they are able to arrive at the sidewall of electrode being etched. The extended shading effect results in negative charges on the electrode being etched [147, 150].

Non-uniformity of plasma potential results in a non-uniform electron current to the wafer being etched, because the electron current exponentially depends on the difference between the plasma potential and the wafer surface potential [146]. The ion current relates with the ion density and the Bohm velocity, which is hardly influenced by the plasma potential [146]. Therefore, the non-uniformity of plasma potential induces unequal local electron and ion currents. A net charging is likely to occur at the wafer surface.

The global potential difference relates to the global non-uniformity of plasma density or spatially imperfect neutralization of the ion beam [165]. During the ion milling process, positive charging is mainly caused by the high energy ion beam and background gas ions; negative charging is mainly caused by the electron shower generated by the neutralizer and plasma electrons [165]. In addition, when high energy ions bombard a material, the electrons might escape the material after receiving enough energy from

Figure 6.6: Illustration of (a) electron shading and (b) extended electron shading.

an energetic ion. So it can be expected that the positive charge on the wafer is even more than the sum of all positive charge implanted. The polarity of net charge may relate with the ion milling parameters, which can be either positive or negative [165, 166].

6.2.4.2 Analysis of the plasma charging in this study

In our experiments, we have found in the virgin I - V curves that the regular PZT capacitors are positively poled while the overetched PZT capacitors are negatively poled. A possible explanation of this observation is given by plasma etching induced charging. Because the second ion milling process uses a smaller ion beam current and a larger neutralizer current than the first ion milling process, it may happen that the four $4 \times 4 \text{ mm}^2$ metal areas are positively charged during the first ion milling process and the $100 \times 100 \text{ }\mu\text{m}^2$ square electrodes are negatively charged during the second ion milling process. The regular PZT capacitors are inside the $4 \times 4 \text{ mm}^2$ areas and are positively poled during the first ion milling. During the second ion milling, the negative charges on $100 \times 100 \text{ }\mu\text{m}^2$ electrodes are compensated by the positive charges on the $4 \times 4 \text{ mm}^2$ metal areas, and do not influence the PZT of the regular PZT capacitors. The overetched PZT capacitors are outside the $4 \times 4 \text{ mm}^2$ metal areas thus are negatively poled during the second ion milling process. Further experiments are needed to study the relation between ion milling parameters and charging polarity in more detail. If the net charge caused by processing is not compensated in time, it may gradually influence PZT properties.

6.3 Plasma etching induced PZT damage

Besides the charging effect, the ion milling process may also directly influence PZT properties and reliability. In this section, we review the reported mechanisms and phenomena of plasma etching induced PZT damage. Subsequently we compare the permittivity, breakdown voltage and breakdown time between the regular and overetched PZT capacitors and give a possible explanation of the measurement results.

6.3.1 Mechanism and phenomenon of etching induced PZT damage

6.3.1.1 Publication review

Plasma etching induced PZT damage has been reported by several authors. The plasma etching process is able to damage PZT material either physically or chemically [136–138, 140, 143]. The physical way stems from the bombardments of energetic ions; the chemical way comes from the contamination of impurities in the plasma. Pan et al. prove that the physical effect of ion bombardments is the main reason of the PZT degradation [138]. Stanishevsky et al. find that the impurities in etching gas contribute to the change of PZT chemical composition [140]. Lim et al. propose that residues like ZrCl and ZrF might worsen electrical properties of PZT [137]. Jung et al. detect the generation of H - O dipoles due to hydrogen reduction during plasma exposure [136]. Soyer et al. show a surface layer developed during plasma etching which contains fluorine and has poor electrical properties [143].

The PZT damage/degradation caused by the physical effect of ion bombardments can manifest itself by the increase of surface roughness, shift of the hysteresis loop, decrease of permittivity, decrease of remnant polarization, and increase of coercive fields [141, 142, 144]. Soyer et al. study the PZT damage as a function of ion milling parameters [144]. The PZT surface roughness increases after etching. This is presumably because the PZT grain boundaries contain more Pb than PZT bulk and the etching rate of Pb oxide is higher than Ti/Zr oxide. As a result, PZT grain boundaries are preferentially etched and the PZT surface roughness increases [144]. The shift of the hysteresis loop indicates the generation of an internal electric field, which may correspond to the charge accumulation or compensation at the top electrode-PZT interface [144]. A damaged layer at the PZT surface (or sidewall) becomes non-ferroelectric due to the ion bombardments. The total capacitance is then the remaining ferroelectric capacitance connected in series (or in parallel, in the case of damaged sidewalls) with the non-

ferroelectric capacitance. Since the non-ferroelectric layer has a much lower permittivity than ferroelectric PZT, the total capacitance (thus the equivalent permittivity) decreases during the ion milling process [141, 142, 144]. The formation of a non-ferroelectric layer causes a reduction of the ferroelectric zone; thus it also explains the reduction of the remnant polarization. The pinning between the damaged layer and the undamaged PZT layer contributes to a loss of mobility of domain walls, resulting in an increase of coercive fields [141, 144].

The aging and recovery of plasma etching induced PZT damage is also studied by some authors. Yang et al. find that the clamping from the substrate produces deep potential wells which prevent part of the spontaneous polarization to switch. The clamping from the substrate is reduced by etching, resulting in an increase of polarization. The damaged PZT sidewall/surface produces shallow potential wells. The domains can be excited out of the shallow potential wells at high electric field. The domain wall motion may worsen the PZT reliability [141]. Some researchers find that the plasma etching induced PZT damage can be partly recovered by annealing [137, 144]. In [137, 139], an increase of leakage current because of plasma etching is detected. [137, 139].

6.3.1.2 Contribution of this study

A pure physical Ar ion milling is chosen for patterning of the PZT capacitors for this study because of the following three reasons. Firstly, in case of pure physical plasma etching (instead of chemical reaction involved plasma etching such as RIE or ICP), the PZT degradation resulting from contamination of impurities (such as ZrCl or ZrF) in the plasma will not happen. In addition, the particle bombardments are assumed to be primarily responsible for the etching [144], in other words, the PZT etching rate of pure physical plasma etching is comparable to chemical reaction involved plasma etching. Finally, by using pure physical ion milling, it is possible to etch a global structure including several materials without having to change the etching gas, which is beneficial for MEMS applications [144]. Therefore, a pure physical Ar ion milling is chosen for patterning of PZT capacitors for this study.

Although the plasma etching induced PZT damage/degradation is clearly observed, no recent publications report on the plasma etching induced PZT reliability issue in detail. In this study, the capacitance difference between regular and overetched PZT capacitors indicates the existence of damaged PZT layer. And we further study how the damaged layer influences PZT reliability: the ramped voltage stress (RVS) breakdown and time depen-

Figure 6.7: Small-signal C - V curves of virgin regular and overetched PZT capacitors.

dent dielectric breakdown (TDDB) of these two kinds of PZT capacitors are compared and explained.

6.3.2 Capacitance comparison

The ion milling process can not only charge the electrode being etched, but also directly influences the PZT permittivity as described above. For small-signal capacitance-voltage (C - V) measurements, a DC bias is applied to the top electrode from +8 V to -8 V then back to +8 V (as indicated by the arrows in Fig. 6.7) with a voltage step of 0.5 V. The capacitance-frequency measurement shows that both kinds of PZT capacitors have a stable capacitance from 5 kHz to 4 MHz. We choose an AC voltage of 30 mV at 20 kHz for the small-signal C - V measurements. For both regular and overetched capacitors, the C - V curves show a butterfly shape. The capacitance peaks appear around the coercive voltages. Pintilie et al. show that the capacitance peaks in the butterfly-shape C - V curves directly relate to switching of ferroelectric polarization caused by the change of DC bias [167]. The asymmetric behavior of the C - V curves, namely the capacitance peak at negative voltage is sharp whereas the one at positive voltage is not, may relate to the asymmetric stacks of the MIM capacitor. The key information from Fig. 6.7 is that the capacitance of the regular PZT capacitor is larger than the overetched PZT capacitor.

Figure 6.8: Histograms of capacitance at DC bias of 0 V.

This capacitance difference between regular and overetched capacitors is clearly shown in the histograms of capacitance at zero DC bias in Fig. 6.8. All the measured regular capacitors are larger than 60 pF; all the measured overetched capacitors are smaller than 45 pF.

A possible explanation of the observed capacitance difference is given below. The damaged layer on the upper surface of PZT in over-etched capacitors is marked by dashed lines in Fig. 6.1. This damaged PZT layer may have lower permittivity than the undamaged PZT layer. The thickness of undamaged PZT is then reduced because of the formation of a damaged PZT layer. Thus the capacitance of overetched PZT capacitors could be greatly decreased. The PZT damage at the sidewall could further lower the capacitance of overetched PZT capacitors. But we assume that the main damage is from the ion bombardments on the PZT top surface, because the area of the etched PZT sidewall is only about $20\text{ nm} \times 100\text{ }\mu\text{m}$ whereas the area of damaged PZT surface is $100\text{ }\mu\text{m} \times 100\text{ }\mu\text{m}$.

The PZT in regular capacitors is protected by the Au metal layer. Although the ion bombardments can also affect PZT in regular capacitors through the influence on the interface between top electrode and PZT, this influence is much less than that caused by direct ion bombardments. Therefore, the capacitance of regular PZT capacitors does not decrease as much as that of the over-etched PZT capacitors.

6.3.3 Reliability comparison

The ramped voltage stress (RVS) measurements of the two kinds of PZT capacitors are compared. In this measurement, the bottom electrode is

Figure 6.9: Ramped voltage stress measurements of both kinds of PZT capacitors.

grounded; a ramped voltage stress from 0 to ± 200 V is applied to the top electrode and the current is measured. The voltage ramp rate is about 0.2 V/s. It slightly varies due to the measured current. The larger the measured current is, the higher is the voltage rise rate [157]. The breakdown voltage is determined from the I - V curve as the point where a sudden current increase occurs. At negative stress voltage, there is no significant difference between these two kinds of PZT capacitors; and the dielectric strength can be larger than 100 MV/m. However, at positive voltage, the regular capacitor has larger dielectric strength (~ 80 MV/m) than the overetched capacitor (~ 40 MV/m), as shown in Fig. 6.9.

The time dependent dielectric breakdown (TDDB) measurements also show this reliability difference between two kinds of capacitors at positive voltage. The Weibull distribution of the TDDB measurement conducted at +70 V is shown in Fig. 6.10. The two kinds of capacitors have similar Weibull shape parameters, which are both smaller than 1. This low value indicates that defects play an important role in the measured PZT breakdown [97]. The regular PZT capacitor has a larger Weibull scale parameter than the overetched PZT capacitor. As described in chapter 2, a larger Weibull scale parameter indicates a longer breakdown time [97], therefore, statistically the regular capacitor breaks later than the overetched capacitor.

Figure 6.10: Time dependent dielectric breakdown measurements on both kinds of PZT capacitors.

To understand the difference between these two kinds of PZT capacitors in the RVS and TDDB measurements, it is necessary to know the difference between their PZT layers. The damaged PZT layer is assumed to be caused by structural defects generated by the ion bombardments [139, 141–144]. Only the PZT in overetched capacitors is directly bombarded by the energetic ions. So those defects are probably present in the upper part of the overetched PZT capacitors, but not in the regular PZT capacitors.

People propose many mechanisms to explain the energetic particle bombardments induced defect generation, such as lattice dislocation, bond disruption, stress modification and plasma species permeation [139]. The positively charged defects, in particular oxygen vacancies, are the most probable ones induced during the ion milling process [139, 141–144]. We speculate that the positively charged defects initiate the dielectric breakdown. Since those defects are generated in the overetched PZT capacitors but not in the regular PZT capacitors, they only contribute to the breakdown of overetched capacitors. Therefore, at positive voltage, the overetched capacitors have lower breakdown voltage and shorter breakdown time than the regular capacitors. At negative voltage, such species are rapidly compensated with electrons and hence do not influence the reliability of PZT underneath. Therefore, we only observe the reliability difference between these two kinds of PZT capacitors in RVS and TDDB measurements at positive voltage.

6.4 Conclusion

This chapter shows that the processing sequence changes the PZT properties and influences the PZT reliability. In this chapter, two kinds of PZT capacitors are compared: the PZT in the overetched capacitors is directly bombarded by the energetic ions during the ion milling process, whereas the PZT in the regular capacitors is protected by the metal layer above. It appears that direct ion bombardments of the PZT surface may cause great PZT degradation/damage. When the PZT is protected by metal, the ion bombardments maybe still cause PZT degradation/damage through modification of the metal-ferroelectric interface, but the PZT degradation/damage is much less than that in case without metal protection.

The plasma charging may influence the poling of the virgin PZT capacitor. The ion milling set-up should be modified to reduce the net charge injected to the electrodes, because the plasma induced charging during ion milling may trigger a dielectric breakdown when the PZT thickness decreases to a certain value or the net charge stays in the electrode for a long time before being compensated. It will be helpful to study the relation between ion milling parameters and plasma charging in more detail, to control the ion milling process induced charging.

The ion milling process probably generates positively charged defects in the upper part of the PZT capacitors when the energetic ions directly bombard on PZT. At negative voltage, such species are rapidly compensated with electrons and hence do not influence the reliability of PZT underneath. At positive voltage, such species could move into the PZT underneath and initiate the dielectric breakdown. Thus applying negative voltage on the top electrode of the PZT capacitors may be one feasible way to reduce the influence of plasma etching induced defects. The oxygen vacancies are possible defects which trigger the breakdown, however, it has not been proved yet. Further study in detail is required to investigate what kinds of defects are generated during ion milling and how they influence the PZT reliability.

Chapter 7

Influence of measurement conditions and sample stacks on PZT reliability

$\text{PbZr}_{1-x}\text{Ti}_x\text{O}_3$ (PZT) exhibits excellent piezoelectric properties, exploited in many MEMS devices, including RF-MEMS switches [134,168]. A prolonged exposure to high voltages is a prerequisite for the application of such devices in products. The reliability of the PZT thin films becomes a decisive factor in the commercialization of piezoelectrically actuated RF-MEMS switches [30,169,170].

Chapter 7 aims to study the physics of PZT degradation and breakdown, by doing ramped voltage stress (RVS) and time-dependent-dielectric-breakdown (TDDB) measurements on unpackaged metal-insulator-metal (MIM) PZT capacitors with various stacks in various measurement conditions. Section 7.1 introduces the MIM PZT capacitors studied in this chapter. Section 7.2 shows that environmental humidity dramatically worsens the PZT reliability according to both RVS and TDDB measurements. Visible breakdown spots on the surface of PZT capacitors are studied in detail. The measurement results in section 7.3 indicate that both reversible and irreversible PZT degradation/breakdown happen during TDDB. In section 7.4, the dependence of time to breakdown on polarity of applied voltage is argued to relate to the PZT crystal structure and the stack of the PZT capacitor. It is also presented how the PZT microstructure, measurement temperature and voltage influence TDDB results.

7.1 Experiments on three kinds of MIM PZT capacitors

In this chapter, we will present not only the influence of measurement conditions but also the effect of electrodes and PZT microstructure on the TDDb behavior of MIM PZT stacks. Therefore, three different kinds of MIM PZT capacitors are prepared for the PZT reliability study in this chapter.

In this study, PZT films are deposited on (111) Pt/Ti/SiO₂/Si substrates using pulsed laser deposition (PLD). To prevent the formation of a pyrochlore phase at the interface between the PZT film and Pt bottom electrode, a thin layer of LaNiO₃ (LNO, 10 nm, deposited by PLD) is inserted between the Pt bottom electrode and PZT film [152], as shown in Fig. 7.1 (a) and (b). A layer of 90-nm-thick Pt is used as the top electrode. 120-nm-thick LNO layers are also used as electrodes of some PZT MIM capacitors under study, as in Fig. 7.1 (c). The PZT layer is 900 nm in Fig. 7.1 (a) and (b); and 350 nm in Fig. 7.1 (c). The top electrode size of all measured PZT capacitors is $100 \times 100 \mu\text{m}^2$. More details of the MIM PZT capacitors fabrication are found in [152].

A Keithley 4200 SCS and a probe station are used for all electrical measurements presented in this chapter. The bottom electrodes of the capacitors are always grounded and the voltages are applied to the top electrodes.

Figure 7.1: SEM cross-section views of MIM PZT capacitors with (a) Pt electrodes and columnar PZT grains, (b) Pt electrodes and dense PZT, (c) LNO electrodes.

The current is measured from the top electrode. The current compliance is set to 100 nA for all TDDB measurements; this compliance level is also used as the breakdown criterion. All the measurements mentioned in this chapter are wafer-level on unpackaged devices.

7.2 Humidity influence on RVS and TDDB measurements of PZT

In this section, the MIM PZT capacitors as in Fig. 7.1 (a) are measured under two different relative humidity (RH) conditions. Both RVS and TDDB results indicate that the environmental humidity greatly worsens the PZT reliability. Some possible explanations to this phenomenon will be given in this section.

7.2.1 RVS measurements and visible damage on the sample surface

RVS measurements (as described in chapter 6) under two RH conditions ($RH > 50\%$ and $RH < 10\%$) are shown in Fig. 7.2. The leakage current of the PZT capacitor increases faster when the humidity is high. During this stress condition, the surface of the PZT capacitor changes over time, namely spots appear as observed using an optical microscope. Under high humidity, these spots occur much more frequently under similar bias conditions as shown in the insets of Fig. 7.2.

These breakdown spots on the top electrode are carefully observed by scanning electron microscopy (SEM), as shown in Fig. 7.3. The breakdown spots in both high and low RH have a similar structure. The outer edge of the breakdown spot is irregular. One (Fig. 7.3 (a)) or several (Fig. 7.3 (c)) holes are observed in the center of the breakdown spot. There are many spherical particles inside the holes. Comparing Fig. 7.3 (b) ($RH < 10\%$) and Fig. 7.3 (c) ($RH > 50\%$), the diameter of the holes in low RH (tens of micrometres) is much larger than that in high RH (several micrometres).

The elemental composition is measured at various positions on the broken surfaces by energy dispersive spectroscopy (EDS). In Fig. 7.3 (c), position *a* outside the breakdown spot is pure Pt. Position *b* between the central hole and the outer edge of the breakdown spot consists of PZT. Position *c*, one of the spheres inside the central hole, is Pt. At position *d* inside the central hole and besides the spheres, we find SiO_x . If we zoom in to see the area marked by the red square in Fig. 7.3 (c), the sidewall of the PZT layer appears as shown in Fig. 7.3 (d). Although inside the

Figure 7.2: RVS measurements under two RH conditions. The insets are optical microscope photos of top electrodes of the measured capacitors at voltages marked by the arrows.

breakdown spot, the PZT columns are still structurally similar to the fresh PZT layer.

In short, we get the following information from SEM/EDS results. At the breakdown spots, the bottom electrode melts and recondensates in the form of spheres on the silicon substrate with an SiO_2 coating. Some PZT have evaporated, leading to holes in the PZT layer at the center of the breakdown spots. The top electrode above and around the PZT holes also melts, resulting in the irregular outer edges of the breakdown spots.

We interpret the appearance of these breakdown spots as the result of local discharges (fed by the capacitive energy of the voltage biased MIM capacitor). The current is not uniformly distributed in the PZT capacitor. We assume that the current concentrates around weak points in the PZT. Given the high overall current that flows during the RVS measurement, the generated Joule heat is able to melt the weak points in the PZT and/or the electrode [171]. Defects are considered to be the main reason of the non-uniformity of current in PZT capacitors [105, 172, 173]. J. S. Lee et al. show that the leakage current through grain boundaries is larger than that through PZT bulk. The increase of grain boundary density leads to a decrease of breakdown field [174]. Waser et al. mention that field-induced local dielectric breakdown is one of the PZT degradation phenomena [173]. The trapped charges may be emitted with the assistance of the electric field [173]. Therefore, it is possible that the defects accumulated at grain

Figure 7.3: SEM micrographs of (a) breakdown spots corresponding to RVS measurement at $RH > 50\%$ and (b) its zoom in image of the red square marked area; (c) one breakdown spot corresponding to RVS measurement at $RH < 10\%$ and (d) its zoom in image of the red square marked area.

boundaries act as the weak points in PZT. The influence of environmental humidity on these weak spots in the PZT material is discussed in section 7.2.3.

7.2.2 TDDB measurements and evolution of the sample surface during TDDB

High environmental humidity also worsens TDDB behavior of MIM PZT capacitors. Fig. 7.4 shows the Weibull distribution of the same type of PZT capacitors on the same wafer (as in Fig. 7.1 (a)) under two humidity conditions. Compared with the TDDB result at low RH, the Weibull scale parameter is significantly decreased at high RH, which indicates a decrease of the breakdown time; the decrease of Weibull shape parameter under

Figure 7.4: Weibull distribution for two RH conditions: the symbol lines are measurement results; the solid or dash lines are linear fitting of the measurement results.

high RH may indicate that high humidity increases the number of (active) defects in PZT and triggers breakdown [97].

According to the optical microscope inspection, the top electrodes of most PZT capacitors do not break during TDDB measurements in low RH. That's why we are able to repeat the TDDB in low RH in section 7.3. However, the top electrodes gradually break during TDDB measurements in high RH, as presented below.

The visible degradation of the top electrode during TDDB at +25 V in RH of 50% is shown in Fig. 7.5. The top electrode begins to break after stress of 30 s. With the increase of time, more and more breakdown spots appear on the top electrode, again indicating that the breakdown spots on the top electrode are defects related. A similar finding is reported in [171].

The corresponding current-time ($I-t$) curve of this TDDB measurement in Fig. 7.6 (a) shows a steady increase followed by sudden death. This TDDB measurement is shortly interrupted several times (as visible in Fig. 7.6 (a)), to measure the capacitance change during the TDDB measurement. The fluctuating increasing leakage current matches the theory of electromigration model of PZT breakdown [111], as described in chapter 2.

The capacitance decreases with time during the TDDB measurement as shown in Fig. 7.6 (b). If we multiply the measured capacitance-voltage ($C-V$) curves by the ratio of the initial capacitance to the decreased capacitance, it will show that the ratio of the capacitance at DC bias of 0 V to the capacitance at DC bias of 10 V does not change, as in Fig. 7.6 (b).

Figure 7.5: Microscope images of top electrode during the TDDB measurement at +25 V in RH of 50%: after (a) 15 minutes, (b) 40 minutes, (c) 70 minutes and (d) 3 hours 40 minutes.

Therefore, we think that the sudden drop of current in the $I-t$ curve at ~ 270 minutes in 7.6 (a) is because the electrode below the probe needle is molten (as observed in optical microscope), disconnecting the electrode and the signal source. The decrease of capacitance in Fig. 7.6 (b) may be explained by the decrease of effective electrode area, such as the partial delamination of the electrode from the PZT, or the melting/evaporation of electrode. The PZT material under unbroken top electrode probably still does not degrade and remains functional.

7.2.3 Analysis of humidity influence on PZT

All the above measurements indicate that high RH results in an increase of (active) defects in PZT. Indeed, when sufficient water is adsorbed and

(a)

(b)

Figure 7.6: (a) Current-time relation of the TDDB measurement at +25 V in RH of 50%; (b) capacitance-voltage curves measured during this TDDB measurement multiplied by the ratio of initial capacitance to the degraded capacitance.

dissociated on the surface of PZT capacitors, significant defects could be generated to cause PZT degradation [175]. Exposure of PZT to water (or NaOH solution) by itself is reported not to degrade PZT [176]. In our experiments, the combined exposure to moisture with a high electric field stress does lead to PZT degradation. Therefore, the extra defects in high RH may relate with the atomic hydrogen (H) generated by electrolysis of water. Pt is known to catalyze the water dissociation to generate atom H [177].

Hydrogen may change PZT properties through incorporation, diffusion

and reaction with PZT. Atomic H can diffuse freely into bulk PZT and as such does not influence PZT properties. However, the penetration of H can induce structural degradation in various ways as described in the publications [178, 179]. Firstly, at PZT grain boundaries or at the electrode/PZT interface, atomic H can react with oxygen ions and generate oxygen vacancies and electrons, resulting in a decrease of resistivity. Secondly, atomic H is indirectly involved in the dissociation of oxygen in the PZT lattice, generating oxygen vacancies. Thirdly, atomic H may form molecular H_2 and induce the reduction of PZT. Metallic Pb is found at grain boundaries and electrode/PZT interface. Pb further diffuses to the electrode interface and causes PZT degradation. Fourthly, the ionization of H in the PZT lattice releases electrons and decreases PZT resistivity. Ionic H^+ could react with oxygen ions in the lattice and form polar hydroxyl bonds H-O, which acts as a fixed dipole and hinders the switching of PZT domains. Finally, molecular H_2 existing at the grain boundaries may lead to cracks/voids [178]. The hydrogen induced PZT degradation seems mainly to happen at grain boundaries according to the mechanisms mentioned above.

7.2.4 Conclusions

In short, environmental humidity reduces the dielectric strength of PZT MIM capacitors. In the application of PZT-based devices, one can avoid the environmental humidity problem by proper encapsulation. Then we can expect identical behavior of the PZT in low RH. For process optimization, we don't want to encapsulate. Then, if we want to measure material quality, it is important to prevent moisture in the measurements. In the remainder of this chapter, we carefully control the RH below 1% during all measurements.

7.3 Repeated TDDB measurements: reversible and irreversible changes in PZT

As described in 7.2.2, some part of the PZT material remains functional even when the top electrode seriously breaks during TDDB in high RH. To study whether the PZT material really breaks after TDDB in low RH, we repeat the (low RH) TDDB measurement in the same condition, and will present the reversible and irreversible changes in PZT in this section.

To eliminate the influence of humidity, the MIM PZT capacitors are heated to 150° C for 10 minutes with dry air flow before the TDDB measurements presented in this section and later sections; we keep the dry air

Figure 7.7: Current-time relation during and between the first and second TDDB measurements of one PZT capacitor at -30 V.

flow on during TDDB measurements to ensure $RH < 1\%$. The MIM PZT capacitor stacks treated in this section are all as in Fig. 7.1 (a).

7.3.1 Current evolution during TDDB and PZT recovery after TDDB

The top electrodes of most PZT capacitors do not break (according to optical microscope inspection) during the first TDDB measurement. After removing the DC stress, the device recovers to the high resistivity state. When the recovery is complete, we repeat the TDDB measurement on the PZT capacitors which have no visible breakdown spots on the top electrodes. Typically, we store the PZT capacitors un-biased for several days between the two TDDB measurements to complete the PZT recovery. The current-time relation of one PZT capacitor during and between the first and second TDDB are shown in Fig. 7.7 (a). The initial stress current at constant voltage during the second TDDB test is even smaller than that during the first TDDB test.

The PZT recovery process is detected by measuring the current after the first 'breakdown' under a DC bias of 5 V (the polarity of this test voltage is the same as the polarity of the stress voltage used for TDDB). The current-time relation of seven capacitors at 5 V is shown in Fig. 7.8. Most of the time, the leakage current decreases with time at 5 V, but with varying decrease rate. For some PZT capacitors, the leakage current even increases at 5 V. The top electrodes of the PZT capacitors mentioned here are not broken during the first TDDB test. When the PZT capacitors stressed to first 'breakdown' are un-biased for a long time, we detect that all of them recover to the high resistivity state.

Figure 7.8: Current revolution of seven PZT capacitors at 5 V after first TDDB measurement.

7.3.2 PZT degradation and breakdown

The Weibull distributions of the two TDDB measurements are shown in Fig. 7.9 (a). The time until second breakdown is always longer than the first TDDB time. We further note that the capacitance of the PZT capacitor decreases after the first TDDB; but does not further change after the second TDDB, as shown in Fig. 7.9 (b).

To understand these measurement results, it is necessary to distinguish PZT degradation from breakdown. PZT degradation is a reversible process because of oxygen vacancy redistribution in PZT and barrier lowering at the electrode/PZT interface under a continuous voltage stress [105]. Usually, the PZT degradation does not change the dielectric properties of PZT, and the PZT film could recover to its initial state when the voltage stress is removed. The mechanism of PZT breakdown is not clear yet; it probably relates to defects in PZT [105]. As reviewed in chapter 2, many mechanisms and models are proposed to explain PZT TDDB behavior, such as the percolation model [93], E model [102], infant mortality model [106] and electromigration model [111].

The current gradually increases in the I - t curve during PZT degradation, whereas the current has a sudden increase to the compliance for PZT breakdown. Usually, people explain the reversible PZT resistance decrease by PZT degradation. The current increase in Fig. 7.10 is a typical I - t curve of PZT degradation corresponding to the theory in [105]. The sample is measured at a relatively low voltage (-20 V). PZT degradation is clearly detected, but no breakdown is observed. The I - t curve in Fig. 7.7

Figure 7.9: (a) the Weibull distribution of two subsequent TDDB measurements of the same PZT capacitors, (b) the capacitance-voltage relation of the fresh PZT capacitor, after the first TDDB, and after the second TDDB.

shows a sudden loss of insulation of PZT material (a typical breakdown behavior), and the PZT resistance decrease is reversible. Therefore, we think that the stress sequence of first TDDB disables some weak links in PZT; both reversible and irreversible PZT degradation/breakdown happen during TDDB.

Figure 7.10: Current-time relation at -20 V constant voltage stress.

7.4 Influence of measurement conditions and sample stacks on PZT reliability

It would be interesting to study how the measurement conditions and sample stacks influence the PZT reliability. In this section, it is shown that the samples of various PZT microstructures have various polarity influence on TDDB; increasing PZT density is beneficial to TDDB behavior; increasing temperature or voltage amplitude shorten breakdown time.

7.4.1 Polarity influence on TDDB and PZT microstructure

When we apply DC voltages with opposite polarities but the same magnitude to the same PZT capacitors at RH<1%, we find a significant and nontrivial polarity dependence of the time to (first) breakdown, as shown in Fig. 7.11.

The dielectric breakdown process in PZT is probably a combination of electronic and thermal breakdown [105]. Since the observed TDDB is reversible, we suppose that the migration of oxygen ions or vacancies contributes to TDDB by forming high conductivity paths for electrons. The continuous flow of electrons may heat PZT or induce trap generation in PZT, and finally triggers breakdown. The time to breakdown may relate with both the time to form the high conductivity path and the leakage current value.

For a PZT capacitor as in Fig. 7.1&7.11 (a), the negative-TDDB time is longer than the positive-TDDB time. When the top electrode is positively stressed, oxygen vacancies may move along the PZT grain boundaries to the bottom electrode, therefore, oxygen ions accumulate underneath the

Figure 7.11: The Weibull distributions of positive/negative TDDDB of three kinds of MIM PZT capacitors corresponding to Fig. 7.1 (a), (b) and (c).

Pt top electrode, resulting in a reduction of the reverse-biased Schottky barrier (the un-doped PZT is p-type [152]). A high conductivity path is formed because of this redistribution of oxygen ions/vacancies. When the top electrode is negatively stressed, the oxygen ions accumulated near the bottom LNO-PZT interface are compensated by the oxygen vacancies in the LNO, so that the reverse-biased Schottky barrier is not reduced. Therefore, at negative voltage, it is relatively difficult to form a high conductivity path and to inject electrons, resulting in a long negative-TDDDB time.

The TDDDB results of PZT capacitors as Fig. 7.1&7.11 (b) are much better, as a higher voltage can be maintained across these capacitors with the same overall dimensions. Figure 7.11 (b) shows that the positive- and negative-TDDDB times are more or less the same. The used materials and dimensions of Fig. 7.1 (a) and (b) are exactly the same. However, the PZT in Fig. 7.1 (a) has a PZT grain size of about 70 nm, and PZT grain columns connect the top and bottom electrodes. Fig. 7.1 (b) has dense PZT with grain sizes larger than 100 nm. Some PZT grain columns look merged together. No PZT grain boundaries directly connect the top and bottom electrode. It is possible that the PZT grain boundaries are the

weakest links. Since no PZT grain boundaries, thus no high conductivity paths, connect the two electrodes in Fig. 7.1 (b), a positive stress on the top electrode does not trigger TDDB easier than negative stress.

As stated, a higher stress voltage is required to observe breakdown in Fig. 7.1 (b) than Fig. 7.1 (a). This is probably because the PZT density in Fig. 7.1 (b) is larger than in Fig. 7.1 (a). Section 7.4.2 describes in detail about the relation between PZT density and breakdown time.

For a PZT capacitor as Fig. 7.1&7.11 (c), the top LNO-PZT interface is rougher than the bottom LNO-PZT interface; and the columnar PZT grains connect the two electrodes. It can be expected that oxygen ions accumulate near the top LNO at positive stress; and accumulate near the bottom LNO at negative stress. Oxygen ions are apparently easier compensated near the rough top electrode than near the bottom electrode. Thus the result is opposite to Fig. 7.1&7.11 (a): positive-TDDB time is longer than negative-TDDB time.

7.4.2 PZT density influence on RVS and TDDB

By changing the PLD parameters, we can get PZT with a microstructure as in Fig. 7.1 (b), which has an even smaller gap between PZT grains and more grains merged together [152]. The RVS measurement results in Fig. 7.12 show that the denser the PZT film is, the larger is the breakdown voltage. Fig. 7.13 shows the TDDB measurements at 45 V. The statistical breakdown time increases with PZT density. From Fig. 7.12 and Fig. 7.13, we can also qualitatively see that both breakdown voltage and breakdown time decrease with temperature. We will give more details of the temperature influence in section 7.4.3.

It has been reported that the reliability improves with an increasing density of perovskite material. Yao et al. point out that the dielectric strength is higher in denser and less rough PLD lead lanthanum zirconate titanate [180]. Uchida et al. introduce a pretreatment for PZT deposition and show that the PZT with higher density has larger breakdown voltage [181]. The mechanism has not been clearly explained yet. It is possible that the increase of PZT density makes the defects accumulation at the grain boundaries more difficult, thus it is harder to trigger the breakdown. Another possible reason is that in the less dense material, there may be more space between the atoms, leading to weaker bonds. If the atoms are bound weaker, the activation energy to release these atoms and let them hop or diffuse through the solid state, will become lower.

Figure 7.12: RVS measurement results of the PZT capacitors as Fig. 7.1 (b) and a denser PZT at temperature of (a) 85 °C and (b) 150 °C.

7.4.3 Influence of measurement temperature and voltage on TDDB results

As reviewed in chapter 2, one purpose of TDDB measurement is to extrapolate a reliable application voltage at which the time to breakdown is longer than 10 years. Temperature acceleration methodology and voltage acceleration methodology are two widely used methods to obtain the reliable application voltage. We have not done the temperature/voltage acceleration for our reliability study yet, however, some of our TDDB measurements are able to show the influence of temperature and voltage on the PZT reliability. In this section, we present the TDDB measurements operated at various temperatures and voltages. The MIM PZT capacitor

(a)

(b)

Figure 7.13: TDDDB measurements at 45 V of the PZT capacitor as Fig. 7.1 (b) and a denser PZT at temperature of (a) 85 °C and (b) 150 °C.

stacks treated in this section are as in Fig. 7.1 (a).

Fig. 7.14 (a) shows the Weibull distributions of TDDDB measurements at various temperatures. We think the distributions are monomodel Weibull type. The Weibull shape parameter (the slope of the Weibull curves in Fig. 7.14 (a)) does not significantly change with temperature. The Weibull scale parameter (the intercept on the y-axis of the Weibull curves in Fig. 7.14 (a)) equals to the time when ~63% of the devices break down. We choose the Weibull scale parameter as the breakdown time and draw the relation between Weibull scale parameter and temperature in Fig. 7.14 (b). The breakdown time is often found to decrease with temperature as

Figure 7.14: (a) TDDDB measurements at various temperatures; (b) relation between the breakdown time and temperature.

described by the Arrhenius law: $t_{BD} = Ae^{\frac{E_a}{kT}}$, resulting in a linear relation: $\ln(t_{BD}) = \ln(A) + \frac{E_a}{kT}$. t_{BD} is the breakdown time; A is a constant; E_a is the activation energy; k is the Boltzmann constant. According to the linear fit in Fig. 7.14 (b), the activation energy is 0.57 eV. Not many data of Arrhenius activation energy can be found in publications. We find one case based on experiments of sol-gel PZT with thickness of 250 nm [182]. The authors find that the Arrhenius activation energy depends on the applied voltage, which are 1.22 eV at 14 V and 1.31 eV at 10 V. It is difficult to compare our result to these values, because of different PZT processing, different PZT thickness and different measurement voltage. It is worth a more comprehensive study with systematically varied PZT thickness and

Figure 7.15: TDDDB measurement results at +30 V and +20 V.

systematically varied stress voltage to find out if this is really Arrhenius behavior, and to study what process governs the PZT degradation. We could not do this study because of the lack of many supposedly identical samples.

Fig. 7.15 shows the Weibull distribution of TDDDB measurement results at various voltages. The Weibull shape parameter at 30 V is smaller than that at 20 V, indicating defects play a more significant role at 30 V [97]. The Weibull scale parameter at 30 V is also smaller than that at 20 V. It matches the common knowledge that the breakdown time becomes shorter at a larger voltage. However, it is hard to say whether our breakdown-voltage relation follows the E model or the $1/E$ model because of the limited statistics. It will be interesting to further study the breakdown time-voltage relation. This study also requires many supposedly identical samples as mentioned above.

7.5 Conclusion

This chapter presents PZT reliability study in various measurement conditions and on various PZT sample stacks.

Humidity greatly worsens PZT degradation and breakdown. Visible changes on the top electrodes are observed during both RVS and TDDB measurements, which are interpreted as the signature of local discharges. Both reversible and irreversible PZT degradation/breakdown are observed to happen during TDDB. Hydrogen related defects may explain the humidity influence on PZT reliability. After a first breakdown and recovery, a PZT capacitor will have a longer TDDB time; be it with a degraded overall capacitance.

At the same composition and layer thickness, the crystal structure of PZT determines the breakdown voltage to a large extent. Denser PZT samples reach higher breakdown voltages in RVS measurements and have longer breakdown time in TDDB measurements. A higher stress temperature or a larger stress voltage leads to a shorter breakdown time, which matches the Arrhenius law and the voltage acceleration knowledge. However, more study is needed to determine an appropriate temperature or voltage extrapolation model.

Chapter 8

Conclusions

This thesis presents characterization and reliability studies towards piezoelectrically actuated RF-MEMS switches. Especially accurate measurements of the switches under test across a wide range in frequency and reliability of piezoelectric thin film actuators have been studied in detail, since these are critical aspects for the project objective of high density piezoelectric RF-MEMS switch arrays to be integrated in an energy-efficient agile RF transceiver with reconfigurable antenna.

In a first step, the research status and performance criteria for RF-MEMS switches are introduced (see chapter 1). The galvanic switches are broadband and have a higher isolation than capacitive switches. The piezoelectric switches have better performance in actuation voltage, speed, linearity than electrostatic switches. So we aim to study towards galvanic RF-MEMS switches using PZT thin film actuators. Reliability, especially the reliability of the PZT thin film actuators, is the main obstacle of development of the target devices.

The main failure modes of RF-MEMS switches, namely stiction, contact resistance degradation, dielectric charging/breakdown, creep, RF-power induced failure, and PZT actuator degradation/breakdown, are summarized in chapter 2. Both mechanical reasons (repeated contact) and electrical reasons (Joule heating, field emission, field evaporation and so on) can result in stiction and contact resistance change. The trade-off between adhesion force and contact resistance should be made in order to choose a suitable material of contact electrodes. The capillary effect and dielectric charging also contribute to stiction. Surface contamination caused by frictional polymers can increase the contact resistance and finally lead to open failure.

The review of dielectric charging/breakdown and PZT reliability in

chapter 2 provides background knowledge of our reliability study on PZT thin films in chapter 6 and chapter 7. The dielectric breakdown mechanism, leakage current, charging and TDDB mechanism are introduced in chapter 2. The TDDB models of PZT thin films (treated in the electrical domain), provide good explanation of the experiments in certain conditions, although the physical nature of PZT TDDB behavior is not clear yet and deserve more studies.

In a second step, an accurate characterization of the switches under test is done to get a more comprehensive understanding of the RF-MEMS switches. A well-designed de-embedding/calibration and an accurate measurement is the basis for reliability study on RF-MEMS switches. The small-signal capacitance of capacitive RF-MEMS switches is measured across 11 orders of magnitude from 10 mHz to 1 GHz. An equivalent circuit, representing the origin of the parasitics, is given. The C - V curves measured by five different methods are compared. The low-frequency and quasi-static C - V curves yield different results from classical high-frequency and radio-frequency C - V curves. The low-frequency and quasi-static capacitance is up to 200% and 25% higher than the high-frequency and radio-frequency capacitance around the pull-in and pull-out voltages (see chapter 3).

A transducer model, which represents the coupling between mechanical and electrical operation of RF MEMS switches, is introduced to explain this observed difference in C - V curves. In the up-state, this can be attributed to the mechanical movement of the top electrode at low-frequency. A one-dimensional spring-mass model quantitatively reproduces the up-state behavior. In the down-state, the difference between low-frequency and high-frequency capacitances is attributed to squeeze-film effects in the residual gas volume between the top electrode and the dielectric film. This assumption is supported by vibrometer measurements, FEM simulations, and C - V measurements at reduced pressure. A concise impedance model for RF-MEMS switches should include these mechanical properties, in particular for biasing voltages close to pull-in and pull-out (see chapter 4).

The measurement of spring constant is important to characterize the RF-MEMS switches; it may be a suitable parameter for in-line process control if a fast and reliable wafer-level electrical measurement is developed. The spring constant of four RF-MEMS capacitive switch designs extracted using vibrometry, pull-in voltage, and low-field C - V measurements, are compared. The vibrometer results deviate from the other estimates. The low-field C - V method is able to easily and quickly detect the change of spring constant and the relation of springs of various designs. However, the absolute value of spring constant obtained from the low-field C - V method

is relatively inaccurate due to measurement noise and parameter uncertainties, which makes this method unsuitable for in-line process control. The spring constant from pull-in voltage is easy to measure and is accurate. The disadvantage is that applying the high pull-in voltage may induce charging. This may not be a serious problem if the high voltage is applied once for the in-line process control. Therefore, the pull-in voltage method is considered a suitable way for in-line process control (see chapter 5).

In a third step, the study goes towards the target piezoelectric switches: the reliability of PZT thin films are investigated, which are potential piezoelectric actuators of the target devices. The processing sequence, especially plasma etching, influences the PZT properties and reliability. By comparing two kinds of PZT capacitors, it appears that direct ion bombardments of the PZT surface may cause strong PZT degradation/damage. The ion milling process induced charging should be controlled, or better still eliminated. The ion milling process probably generates positively charged defects in the upper part of the PZT capacitor when the energetic ions directly bombard on PZT. The reliability is significantly worsened when positive voltage is applied on the top electrode, whereas not much change in reliability is observed by applying negative voltage to the top electrode (see chapter 6). Further study is required to investigate what defects are generated during ion milling and how they influence the PZT reliability.

In addition, the measurement conditions and stacks of PZT MIM capacitors also influence the reliability. Humidity greatly worsens PZT degradation and breakdown. To measure PZT material quality, it is important to prevent moisture in the measurement. Both reversible and irreversible PZT degradation/breakdown are observed to happen during TDDB. At the same composition and layer thickness, the crystal structure of PZT determines the breakdown voltage to a large extent. A higher temperature or a larger voltage leads to a shorter breakdown time (see chapter 7). More study is needed to determine an appropriate voltage/temperature extrapolation model to be used for reliability estimates of PZT-MIM based MEMS devices.

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Summary

This PhD research is part of the EPAMO project aiming to develop high density piezoelectric RF-MEMS switch arrays to be integrated in an energy-efficient agile RF transceiver with reconfigurable antenna. According to the comparison of performance criteria for various RF-MEMS switches, the galvanic switches are broadband and have a higher isolation than capacitive switches. The piezoelectric switches have better performance in actuation voltage, speed, linearity than electrostatic switches. So we aim to study towards galvanic RF-MEMS switches using PZT thin film actuators. This thesis presents characterization and reliability studies towards piezoelectrically actuated RF-MEMS switches.

The reliability problem, especially the reliability of the PZT thin film actuators, is the main obstacle of the development of the target device. The main failure modes of RF-MEMS switches include stiction, contact degradation, dielectric charging/breakdown, creep, RF power induced failure, and PZT actuator degradation/breakdown. The dielectric breakdown mechanism, leakage current, charging and TDDB mechanism are introduced in chapter 2 and provide a background knowledge of the reliability studies on PZT thin films in chapter 6 and 7.

An accurate characterization of the switches under test across wide range in frequency is important not only for a comprehensive understanding of the device but also for detecting the degradation/breakdown of the device. Chapter 3 focuses on the calibration and de-embedding of pF-level capacitance measurements, and provides an equivalent circuit presenting the origins of all the parasitics. The small-signal capacitance of the switches under test is measured across 11 orders of magnitude in frequency from 10 mHz to 1 GHz. The capacitance-voltage (C - V) curves measured by five different methods are compared. The low-frequency and quasi-static C - V curves yield different results from classical high-frequency and radio-frequency C - V curves. The low-frequency and quasi-static capacitance is up to 200% and 25% higher than the high-frequency and radio-frequency capacitance around the pull-in and pull-out voltages. This phenomenon

is explained by a transducer model expressed in chapter 4. A transducer model presents the coupling between mechanical and electrical operations of RF-MEMS switches. In the up-state, a one-dimensional spring-mass model quantitatively reproduces the C - V behavior. In the down-state, the difference between low-frequency and high-frequency capacitances is attributed to squeeze-film effects in the residual gas volume between the top electrode and the dielectric film. This assumption is supported by measurements at reduced pressure.

The measurement of spring constant is important to characterize the RF-MEMS switches; it may be a suitable parameter for in-line process control if a fast and reliable wafer-level electrical measurement is developed. The spring constant of four RF-MEMS capacitive switch designs extracted using vibrometry, pull-in voltage, and low-field C - V measurements, are compared in chapter 5. The pull-in voltage method is considered a suitable way for in-line process control.

In chapter 6 and 7, the study goes towards the reliability of metal-insulator-metal (MIM) capacitors with PZT thin film, which are potential piezoelectric actuators of the target device. The influence of plasma etching process on PZT properties and reliability are studied in chapter 6. The ion milling process induced charging should be controlled, or better still eliminated. By comparing two kinds of PZT capacitors, it appears that direct ion bombardments of the PZT surface may cause PZT degradation/damage: the reliability is significantly worsened when positive voltage is applied on the top electrode, whereas not much change in reliability is observed by applying negative voltage to the top electrode.

Chapter 7 studies the influence of the measurement conditions and the stacks of PZT MIM capacitors on the reliability. Humidity greatly worsens PZT degradation and breakdown. To measure PZT material quality, it is important to prevent moisture in the measurement. Both reversible and irreversible PZT degradation/breakdown are observed to happen during TDDB. At the same composition and layer thickness, the crystal structure of PZT determines the breakdown voltage to a large extent. A higher temperature or a larger voltage leads to a shorter breakdown time.

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List of Publications

Journal Papers

1. **Jiahui Wang**, Evert Houwman, Cora Salm, Minh Nguyen, Kurt Vergeer, Jurriaan Schmitz, ‘Process induced poling and plasma induced damage of thin film PZT’, *Microelectronic Engineering*, Volume 177, 2017, PP: 13-18.
2. **Jiahui Wang**, Jeroen Bielen, Cora Salm, Gijs Krijnen, Jurriaan Schmitz, ‘On the capacitance of RF MEMS switches at very low frequencies’, *IEEE Journal of the Electron Devices Society*, Volume 4, Issue 6, 2016, PP: 459-465.
3. **J. H. Wang**, J. Zhou, W. L. Zhou, H. Tong, D. Q. Huang, J. J. Sun, L. Zhang, X. M. Long, Y. Chen, L. W. Qu, X. S. Miao, ‘A high speed asymmetric T-shape cell in NMOS-selected phase change memory chip’, *Solid State Electronics*, Volume 81, 2013, PP: 157-162.
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Conference Proceedings

1. **Jiahui Wang**, Cora Salm, Evert Houwman, Minh Nguyen and Jurriaan Schmitz, ‘Humidity and polarity influence on MIM PZT capacitor degradation and breakdown’, *IEEE International Integrated Reliability Workshop (IIRW)*, South Lake Tahoe, CA, USA, 2016, PP: 65-68.
2. **Jiahui Wang**, Jeroen Bielen, Cora Salm, Jurriaan Schmitz, ‘Spring-constant measurement methods for RF-MEMS capacitive switches’, *IEEE International Conference on Microelectronic Test Structures (ICMTS)*, Yokohama, Japan, 2016, PP: 10-14.
3. **Jiahui Wang**, Cora Salm, Jurriaan Schmitz, ‘Comparison of C-V measurement methods for RF-MEMS capacitive switches’, *IEEE International Conference on Microelectronic Test Structures (ICMTS)*, Osaka, Japan, 2013, PP: 53-58.

Local Workshop

1. **Jiahui Wang**, Cora Salm and Jurriaan Schmitz, ‘Comprehensive C - V measurement and model for RF-MEMS capacitive switch from DC to RF’, Workshop on Semiconductor Advances for Future Electronics (SAFE), Amersfoort, The Netherlands, 2016.
2. **Jiahui Wang**, Cora Salm and Jurriaan Schmitz, ‘Reliability of PZT capacitors: TDDB, RVS and degradation’, MESA+ yearly workshop, Enschede, The Netherlands, 2014.
3. **Jiahui Wang**, Cora Salm and Jurriaan Schmitz, ‘Overview of galvanic RF-MEMS switches’, Workshop on Semiconductor Advances for Future Electronics (SAFE), Rotterdam, The Netherlands, 2012.