Geometry-Dependent Quality Factors in Ba$_{0.5}$Sr$_{0.5}$TiO$_3$ Parallel-Plate Capacitors

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Abstract—Quality factor variation with top electrode area in thin-film barium–strontium–titanate parallel-plate capacitors is discussed. At low frequencies, the geometry dependence is consistent with the presence of a parallel parasitic loss pathway enabled by conduction over the device mesa surface. At high frequencies, the variation in the quality factor with top electrode area is due to a geometry-independent series resistance.

Index Terms—Geometry scaling, high-$k$, parallel-plate capacitor, quality factor.

I. INTRODUCTION

Over the past two decades, significant progress has been made in the development of paraelectric Ba$_x$Sr$_{1-x}$TiO$_3$ barium–strontium–titanate (BST) thin films for RF/microwave applications [1]–[8]. Efforts in materials development have led to materials with lower losses [7], [9] and higher tunabilities [10]. At the same time, these materials have been successfully integrated into both passive and active [11] monolithic microwave integrated circuits (MMICs), allowing for the realization of compact, low cost, functional oxide electronics.

One puzzling observation that we have encountered with our parallel-plate BST capacitors is that quality factor $Q$ varies as a function of device area. At low frequencies, the quality factor of large area devices is nearly constant. The $45 \times 45$ $\mu$m$^2$ device in Fig. 1 exhibits this behavior. As top electrode area decreases, the magnitude of $Q$ decreases, and the frequency dependence of $Q$ increases. At microwave frequencies, as shown in Fig. 2, the quality factors for all sizes of devices are grossly frequency dependent, and smaller area devices have higher $Q$. Area-dependent quality factors have also been observed in similarly sized cubic pyrochlore bismuth–zinc–niobate parallel-plate capacitors [12], [13]. Thus, we believe that this behavior may be common to other small-geometry high-permittivity devices. In this paper, we discuss the possible role of extrinsic factors in BST parallel-plate capacitor characterization, and present simple models for the observed geometry dependence of $Q$ in these devices. To that end, we will begin with a description of the devices used in our experiments and their generic equivalent circuit, followed by a discussion of low-frequency properties, and finally high-frequency properties.
A. Device Structure and Fabrication

A simple stacked parallel-plate device structure was used for these measurements. This structure was chosen because it has a well-defined capacitor area and is easy to fabricate. This simple structure also has lower electrode resistances and inductances when compared to standard microwave parallel-plate capacitors [14]. These devices are probed directly from above with coplanar probes. This constrains the top electrode to be at least as large as the signal probe footprint. The probes used in our experiments were the Cascade Microtech ACP40-GSG-50, with 50-μm pitch (probe tip spacing) and a signal probe footprint of 30×30 μm, and 140-GSG-100, with 100-μm² pitch and a signal probe footprint of 12×12 μm².

The devices were fabricated using a two-layer contact lithography process. BST films were deposited by RF-magnetron sputtering on to platinized c-plane sapphire substrates. The device mesas were defined by an etch in undiluted buffered hydrofluoric acid. Platinum top electrode contacts, along with contacts to the ground plane, were then created using a lift-off procedure. The platinum top and bottom electrodes were 150- and 200-nm thick, respectively. The BST film was determined to be 135-nm thick by surface profilometry. Details regarding the BST growth conditions can be found in [10].

B. Equivalent Circuit

The equivalent circuit for a parallel-plate capacitor is shown in Fig. 3. In this circuit, the impedances contains both intrinsic and extrinsic contributions. Specifically, the series resistance and inductance (Rₛ and Lₛ) include both the intrinsic properties of the metal electrodes, as well as the contact resistances from the metal-film interfaces and inductance due to the electrodes. The capacitance and parallel resistance (C, Rₚ) include both intrinsic film properties along with any parallel parasitics such as fringing capacitance, inductance due to the electrical length of the device, and any contributions relating to the device mesa.

The total impedance of the device is

\[ Z = jωLₛ + Rₛ + \frac{Rₚ(1 - jωRₚC)}{1 + ω²Rₚ²C²}. \]  

It should be noted that the terms in (1), particularly C and Rₚ, are all potentially frequency dependent. In particular, in our devices, the ac resistance contribution to Rₚ behaves as an imaginary capacitance, following a fractional power law similar to the capacitance [15]–[17]. At low frequencies, when the impedance of the capacitor is high, the electrode contributions are negligible, and the equivalent circuit can be approximated by a parallel resistor-capacitor (Rₚ-C) combination. Thus, the low-frequency Q, dominated by the parallel resistance, is given by

\[ Q = ωRₚC. \]  

Rₚ contains contributions from both the film’s dc and ac resistances. At very low frequencies, where the dc resistance dominates, Rₚ is roughly constant. At higher frequencies where the film’s ac resistance dominates, Rₚ varies as ω⁻ⁿ, where n ~ 1. This results in a weakly frequency-dependent Q.

At high frequencies, the impedance of the capacitor is small so the equivalent circuit can be approximated as a series resistor-inductor-capacitor (Rₛ-Lₛ-C) combination, making the high-frequency Q dominated by the series resistance. After factoring out the series inductance, the high-frequency Q becomes

\[ Q = \frac{1}{ωRₛC}. \]  

where Rₛ has a weak frequency dependence. Thus, the variation in Q as a function of frequency can be illustrated as shown in Fig. 4.

II. LOW FREQUENCY

Parallel-plate devices are often characterized at low frequencies (1 MHz) because the device properties are closer to the film properties. It is important to note that this does not imply that measured low-frequency device properties are necessarily intrinsic film properties; contributions of parallel parasitic pathways are still important at low frequencies. In practice, the room-temperature behavior of our BST capacitors is consistent with universal relaxation [15]–[17], which means that the measured device properties are not dominated by intrinsic film properties [18]. The dependence on the top electrode area of Q in Fig. 1 is a clear example of how extrinsic factors can influence device properties at low frequencies. At 1 MHz, the devices with larger electrode areas have larger Q, but at 100 MHz, the trend is reversed.

To further investigate this low-frequency geometry dependence, a set of devices with different top electrode sizes and different mesa ledge sizes were fabricated, as shown in Fig. 5(b). On the mask used for fabrication, the top electrode

![Fig. 3. Equivalent circuit for a parallel-plate capacitor.](image)

![Fig. 4. Frequency dependence of Q showing very low-frequency, low-frequency, and high-frequency behaviors.](image)
areas $A$ ranged from $12 \times 12 \, \mu m^2$ to $45 \times 45 \, \mu m^2$, while the mesa ledge widths $L_m$, as indicated in Fig. 5(a), ranged from 5 \, \mu m to 15 \, \mu m. Electrode and mesa lengths and widths were measured using an optical microscope with the photomask features as a length standard. Electrical measurements on these devices revealed that $Q$ was not only a function of top electrode area, but also possibly of the device mesa ledge width as well, as summarized in Fig. 6.

The measured device admittance $Y_{tot}$ can be expressed as the sum of an area-dependent term $Y_A$ and a perimeter-dependent term $Y_P$ as follows:

$$Y_{tot} = Y_A + Y_P.$$  

(4)

We would like to characterize the parallel-plate capacitance and conductance, which are included in $Y_A$, as follows:

$$Y_A = g_d A + j\omega c A.$$  

(5)

where $A$ is the top electrode area, $\omega$ is the radian frequency, $g_d$ is the conductance density, and $c$ is the capacitance density. The quality factor associated with $Y_A$ is clearly geometry independent; the presence of geometry dependence in the $Q$ associated with $Y_{tot}$ indicates a nonzero $Y_P$ contribution.

A variety of factors contribute to $Y_P$. One well-known contribution is fringing capacitance. Analysis of fringing contributions from BST tunability curves reported in [19] revealed a fringing capacitance $C_f$ of the form

$$C_f = \kappa \frac{P}{d}.$$  

(6)

where $P$ is the top electrode perimeter, $d$ is the film thickness, and $\kappa = 0.6 \, fF$ for $d = 210 \, nm$. It should be noted that such analysis may pick up other contributions to $\text{Im}(Y_P)$ that are not tunable by an applied dc bias. The geometry variation in $Q$ resulting from the addition of a fringing capacitance

$$Q = \frac{\omega c A + \omega \kappa P}{g_d A}.$$  

(7)

causes $Q$ to increase as the ratio of perimeter to area $(P/A)$ increases. This is the opposite of the trend observed in Fig. 6.

The observed geometry variation in $Q$ is consistent with the addition of a perimeter-dependent conductance term. A nonzero surface conductivity over the path depicted in Fig. 7(b) enables a distributed perimeter-dependent conductance $G_P$, which appears in parallel with the film conductance $g_d A$, where $G$ is a linear conductance density. $P$ is the top electrode perimeter, $g_d$ is the film conductance density, and $A$ is the top electrode area. It should be noted that $G$ is a function of $L_m$; the functional form of this relationship depends on the surface and bulk material parameters, as will be explained below. With the addition of this conductance term, $Q$ can be expressed as

$$Q = \left(\frac{\omega c}{g_d}\right) \left(1 + \frac{G}{g_d A} \frac{P}{d}\right)^{-1}.$$  

(8)
Fig. 8. 1-MHz $Q$ as a function of the top electrode perimeter to area ratio ($P/A$) for five different mesa ledge widths $L_m$. Measured data points are indicated by solid dots, while fits to (8) are shown as dashed curves.

Since both the capacitance $cA$ and film conductance $gdA$ scale with $A$, the influence of the perimeter-dependent conductance term scales with the ratio of perimeter-to-area $P/A$. With square electrodes, as $A$ decreases, $P/A$ increases. Fig. 8 shows the results of data fits at 1 MHz. The corresponding conductance values and related quantities are listed in Table I. Leakage current measurements failed to reveal any top electrode geometry dependence, as shown in Fig. 9. Thus, dc leakage over/through the device mesa does not contribute to $G$. Comparing the results for devices with different mesa sizes, we find that the values of $G$ saturate for the larger device mesas.

This behavior is consistent with a distributed parasitic pathway. Fig. 10 illustrates an incremental length $\Delta z$ of a transmission line model for the parasitic pathway. The admittance $Y_P$ of the parasitic pathway approaches the characteristic admittance $Y_0$ of the line for large $L_m$. The characteristic admittance of this structure is

$$Y_0 = \left[ G + j\omega C \right] / R$$  

(9)
where $R$, $C$, and $G$ are the incremental resistance, capacitance, and conductance of the parasitic conduction pathway, respectively. The propagation constant $\gamma$ for the line is

$$\gamma = \sqrt{R(G + j\omega C)}.$$  \hspace{1cm} (10)

For our pathway, $C$ and $G$ are proportional to $P$, while $R$ is inversely proportional to $P$. Therefore, $Y_0$ is proportional to $P$ and $\gamma$ is independent of $P$. Our pathway can be modeled as a short-circuit terminated transmission line of length $L_m$. Thus, the input admittance of the parasitic pathway is

$$Y_p = Y_0 \left[ \tanh(\gamma L_m) \right]^{-1}.$$  \hspace{1cm} (11)

The conductance correction is

$$G_P = \text{Re}\{Y_0\} \text{Re}[\left[ \tanh(\gamma L_m) \right]^{-1}]$$

and the capacitance correction is

$$\frac{\text{Im}\{Y_p\}}{\omega} = \frac{\text{Re}\{Y_0\}}{\omega} \text{Im}[\left[ \tanh(\gamma L_m) \right]^{-1}]$$

and

$$\frac{G_P}{\text{Re}\{Y_0\}} \approx \frac{\text{Im}\{Y_p\}}{\omega}.$$  \hspace{1cm} (15)

Thus to determine the characteristic admittance of the parasitic pathway, both capacitance and $Q$ must be examined. Fig. 11 shows the variation in capacitance density for devices with different top electrode areas and mesa ledge widths. The capacitance density decreases with smaller top electrode areas (larger $P/A$). While $Y_p$ always includes a capacitance correction, the correction should cause the capacitance to increase with larger $P/A$. This is the opposite of the trend observed in the measured data. Fringing would also result in an increase with larger $P/A$. This behavior cannot be explained by a parasitic inductance because such a parasitic inductance, either in series or parallel, would need to be unphysically large (i.e., $> 100 \mu$H).

Another explanation for the observed decrease in capacitance density for larger $P/A$ could be a systematic error in the measurement of the fabricated electrode areas. Since the top electrodes are square, an offset in width/length measurements would produce a term proportional to perimeter as a correction to the top electrode area. The sign of this term is negative if the measured length/width is smaller than the actual length/width. For the data in Fig. 11, an error of approximately 1 $\mu$m changes the sign of the perimeter-dependent term, causing capacitance density to increase with larger $P/A$.

The photomask features used as a length standard were fabricated with a $\pm 0.25-\mu$m tolerance. In the photographs used to compare the mask features with the fabricated devices 1 $\mu$m (i.e., 1/12 of a 12-\mu feature) was easily resolvable. The measured electrode lengths/widths were typically 0.5 $\mu$m larger than the mask features. This is consistent with the image-reversal process used to define the electrodes, which tends to result in larger features than the mask, rather than smaller features, which a 1-$\mu$m error would require. Thus, it is unlikely that error in length/width measurements alone are the cause of the capacitance density variation. It is important to note that this does not change the geometry-dependent behavior of $Q$ observed in Fig. 8.

III. High Frequency

The simplified high-frequency equivalent circuit for a parallel-plate capacitor consists of the series combination of $R_s$, $L_s$, and $C$. The measured self-resonance frequencies, between 4 and 16 GHz for these devices, were used to approximate $L_s$. The measured reactance was then corrected for this contribution, leaving the capacitance. The associated quality factors for these devices are shown in Fig. 2. The behavior of capacitance frequencies is dominated by series resistance, agreeing with (3). The geometry dependence at high frequency is evident in Fig. 2 where smaller area devices have higher $Q$.

If both the capacitance and series resistance were proportional to area, $Q$ would be geometry independent. The presence of a geometry-independent series resistance changes this behavior, giving smaller area devices a higher $Q$ because of their larger reactance. Fig. 12 shows how $Q$ is expected to vary for devices with different areas, but the same series resistance. This is consistent with the behavior in Fig. 2.

Comparison of Figs. 2 and 12 reveals that the geometry-independent series resistance component in the measured data

![Graph](image-url)
has some frequency dependence. This is evident when the measured data is compared to the $1/f$ reference line; the traces in Fig. 2 deviate from $1/f$, while the traces in Fig. 12 do not. Thus, while the addition of a geometry-independent frequency-independent series resistance qualitatively describes the geometry dependence of the high-frequency quality factors, a better model would include some frequency dependence in the series resistance term.

Two geometry-independent sources of series resistance can be identified for the parallel-plate structures under consideration: the bottom electrode and the coplanar probe. The bottom electrode resistance is largely determined by the probe pitch, the lateral spacing between the centers of the top electrode and ground plane contacts. This resistance is in-plane and, therefore, inversely proportional to the bottom electrode thickness. When the device is small compared to the probe pitch (spacing between probe feet)—which is usually the case for the small devices used for high-frequency measurements, but not necessarily the case for large low-frequency devices—the series electrode resistance is easily overwhelmed by this nongeometry-dependent contribution. Another source of geometry-independent series resistance is the coplanar probe used for characterization. In principle, the combination of calibration and the measurement of shorted structures (as in Fig. 13) should remove these contributions from the measured device properties, but there are reasons why this may not always be the case.

Series resistances are typically evaluated using a shorted device structure. Our shorted device structure consists of top metal contacts deposited directly upon the ground plane, as shown in Fig. 13. While this structure is literally the parallel-plate structure without a dielectric, it fails to capture all of the series resistances. Fig. 14 illustrates how a series resistance component remains after deembedding a shorted structure. The weakly frequency-dependent calculated $Q$ values in Fig. 14 were obtained by fitting measured broadband capacitance relaxation to a fractional power law. The details and applicability of this technique are discussed elsewhere [15], [17]. The behavior of the calculated $Q$ is similar to that of the low-frequency $Q$ because the capacitance relaxation in both frequency ranges obeys the same fractional power law [15], [16]. The weak frequency dependence in the calculated $Q$ corresponds to the weak frequency dependence in the capacitance (which has been corrected for $L_{sh}$). The presence of a series resistance causes $Q$ to decrease strongly with frequency. The deembedding procedure using a shorted device fails to fully account for the series resistance of the device; the measured shorted resistance $R_{sh}$ of approximately 0.4 $\Omega$ is smaller than the $\sim 1$ $\Omega$ resistance suggested by the data. This
results in a deembedded \( Q \) that continues to decrease as \( 1/\omega \).

This failure also leads to geometry-dependent deembedded \( Q \) factors.

To understand why the shorted device fails to fully capture the series resistances, we must consider the differences between the device-under-test and the corresponding shorted structure [20].

Two differences between these structures are the contributions of the electrode and the contact resistance at the electrode-film interface to the series resistance. The contact resistance is completely neglected in the shorted device structure, while the electrode resistances are only partially accounted for. The spreading resistance in the top contact and directly below the device are different in the shorted structure, as illustrated in Fig. 13.

The electrode resistances in the device, shown in Fig. 13(a), can be expressed as

\[
\left( \frac{R_s}{L} \right)_{\text{device}} = \frac{1}{2} \left( \frac{\rho}{l_b} + \frac{1}{3} \frac{W}{2} \frac{\rho}{l_b + \frac{1}{2} t_e} + \frac{1}{3} \frac{W}{2} \frac{\rho}{l_b + \frac{1}{2} t_e} \right)
\]  

(17)

where \( \xi \) refers to the lateral spacing between the top and side contacts, \( L \) and \( W \) refer to the lengths and widths of the top and side contacts, \( t_e \) is the top metallization layer thickness, \( l_b \) is the bottom metallization layer thickness, and \( \rho \) is the resistivity of platinum. The prefactor of 1/2 appears because of the symmetry of the coplanar ground–signal–ground probe. The factors of 1/3 appear because of the distributed nature of the spreading resistance [14]. The resistances in the shorted structure, shown in Fig. 13(b), can be expressed as

\[
\left( \frac{R_s}{L} \right)_{\text{short}} = \frac{1}{2} \left( \frac{\rho}{l_b} + \frac{1}{3} \frac{W}{2} \frac{\rho}{l_b + t_e} + \frac{1}{3} \frac{W}{2} \frac{\rho}{l_b + t_e} \right) 
\]  

(18)

Comparing the device with the shorted structure, we find that the spreading resistance of the top contact has combined with the spreading resistance in the ground plane beneath the device (the second and third terms in (17), respectively), yielding a spreading resistance through the stack of both metallization layers [the second term in (18)]. If \( t_b = t_e \), these spreading resistances in the shorted structure are one-quarter of those in the actual device.

We have established that the spreading resistance in the top electrode and directly beneath the device are not properly accounted for by the shorted device structure. However, this failure will not result in geometry dependence because the active area of the device scales with the top electrode area. Similarly, the resistance neglected from the electrode-film interface should also scale with the top electrode area, and is, therefore, not a source of geometry-independent series resistance either.

For a geometry-independent resistances we must look to the lateral resistance in the ground plane and the coplanar probe used for measurements. For device mesas much smaller than the probe pitch, the lateral resistance contribution becomes independent of device size. We do not believe that this lateral ground-plane resistance is the source of our geometry-independent resistance because, for our device geometries, 12 \( \times \) 12 \( \mu \)m

\[2\]  

to 45 \( \times \) 45 \( \mu \)m \( ^2 \) 100 \( \mu \)m pitch devices with 25 \( \times \) 25 \( \mu \)m \( ^2 \) to 45 \( \times \) 45 \( \mu \)m \( ^2 \) ground-plane contacts, the lateral spacing variation, of \( \xi = 55 \) to 81.5 \( \mu \)m, should yield a geometry-dependent resistance. In our measurements, we find the series resistance of all of the devices in this geometry series to be approximately 1 \( \Omega \). Additionally, in [13], bismuth–zinc–niobate parallel-plate capacitors fabricated on gold and platinum bottom electrodes were compared. The gold bottom electrodes were found to improve device \( Q \), but not affect the additional series resistance term. This suggests that something completely independent of device size and geometry, such as the coplanar probe, rather than the electrode metallization, dominates the device’s series resistance.

The coplanar probes used for measurement may also contribute to series resistance; we observe higher series resistance values when using probes with smaller footprints. In particular, we find that the series resistance measured using Cascade Microtech’s I40-GSG-100 probe, with a 12 \( \times \) 12 \( \mu \)m \( ^2 \) footprint, is typically 1.0 \( \Omega \) regardless of device size. Using an ACP40-GSG-50 probe, with a larger 30 \( \times \) 30 \( \mu \)m \( ^2 \) footprint, the series resistance tends to be around 0.15 \( \Omega \). The devices probed with each probe consistently had similar series resistances, regardless of device geometry. The ratio of resistances, i.e., 6.67, is very close to the ratio of the different probe areas, i.e., 6.25.

If this resistance results from physical contact with the top electrode, it will not be corrected during instrument calibration. While the shorted structure may be expected to account for this resistance, our devices are more similar to open-circuit loads than short-circuit loads, limiting the ability of the deembedding procedure using a shorted structure alone to fully capture the electrode-related properties [20]. It should be noted that, in conventional high-frequency parallel-plate capacitor designs, where the devices are not probed directly from above, an open structure is used along with the shorted structure for deembedding. The fact that the series resistances in our devices are so large supports the idea that a series resistance from the coplanar probe contributes to our series resistance.

IV. CONCLUSION

We have shown that geometry-dependent parallel-plate BST capacitor quality factors can be attributed to the influence of extrinsic loss contributions. At low frequencies, we observe that smaller devices have lower \( Q \), and we have found this to be consistent with a parallel parasitic pathway enabled by surface conduction over the unpassivated device mesa. Experiments involving passivation of the exposed mesa surface are currently underway. At high frequencies, we have found the geometry dependence to be the result of a geometry-independent series resistance. This series resistance is larger than what would normally be associated with the device electrodes, even after taking into account that electrode resistances are not fully deembedded using a shorted device alone. The series resistance appears to be related to the choice of coplanar probe used for measurements. Remnants of this series resistance remain in deembedded data.

1The frequency dependence of the measured quality factors in Fig. 14 is weaker than \( 1/\omega \) because of the influence of \( R_c \); this frequency range is the crossover between (2) and (3).

2The additional series resistance term in [13] was incorrectly assumed to be geometry dependent based on the observation of a geometry-dependent \( Q \).

3The devices probed with both probes had the same metallization thicknesses for both the ground plane and top electrode.
resulting in a Q with geometry dependence, as well as a $1/\omega$ frequency dependence.

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REFERENCES


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