A High- Q Widely-Tunable Gigahertz Electromagnetic Cavity Resonator

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Abstract

This paper describes the design, fabrication and testing of a quasistatic electromagnetic cavity resonator fabricated using KOH etching, shallow RIE, metalization and wafer bonding. The resonator is distinguished by its simultaneous high Q near 200, and wide high-frequency tuning range, 2.5 GHz to 4.0 GHz for the experimental resonator presented here. When combined with an integrated actuator, it should be suitable for use in electronically-tunable RF band-pass filters and oscillators. The experimental resonator, however, is tuned with an external piezoelectric actuator for simplicity.

1. Introduction

RF systems such as cell phones and local wireless communication networks often need high-frequency widely-tunable high- Q band-pass filters for channel selection filters and local oscillators. This paper describes the design, fabrication and testing of a quasistatic electromagnetic cavity resonator designed for such applications. It is fabricated using standard MEMS technologies, and is distinguished by its simultaneous high Q near 200 or higher, and its wide high-frequency tuning range, 2.5 GHz to 4.0 GHz for the experimental resonator presented here. While the experimental resonator is tuned with an external piezoelectric actuator, it could be tuned with integrated internal actuation.

Figure 1 shows a cross section of the resonator. It consists of a square central capacitor surrounded by a square toroidal inductor etched jointly as a single cavity into a bonded pair of silicon wafers. The cavity is lined with gold to enable high- Q operation. Both electric and magnetic coupling to the resonator are possible. The figure shows magnetic coupling using a wire loop inserted through the wall of the cavity. Alternatively, electric coupling could be made through direct connections to the capacitor electrodes. Because the cavity is smaller than an electromagnetic wavelength, it is

Figure 1: cross section view of a cavity resonator with magnetic coupling.

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quasistatic with identifiable capacitor and inductor regions. This simplifies tuning, which is performed by compressing the cavity to close the capacitor gap.

Alternative technologies provide wide tuning or high Q , but not both. For example, crystal resonators exhibit exceptionally high Q , but they are generally not tunable. Mechanical MEMS resonators can also exhibit very high Q , but they are not widely tunable, and do not generally operate in the gigahertz range [1,2,3,4,5]. Finally, resonators that employ tunable MEMS capacitors can be widely tunable [6], but the associated inductor, often formed through surface micromachining or with a bonding wire, is usually so lossy that a high- Q resonator can not be obtained [7,8,9]. In contrast, the cavity inductor shown in Figure 1 exhibits a higher Q .

2. Fabrication

Figure 2 shows the resonator fabrication process flow. A silicon wafer is first masked using silicon-nitride (Si_3N_4), and then etched using potassium-hydroxide (KOH) to form the inductor cavity (Figure 2A). Corner compensation [10] is used to preserve the shape of an interior square island that will support the capacitor electrode. A single-sided etch is used to form a half cavity, while a double-sided etch is used to form an additional slit through which a magnetic coupling loop may be inserted. Next, reactive-ion etching is used to remove the Si_3N_4 mask over the capacitor island, and additional KOH etching is used to open the capacitor gap (Figure 2B). The KOH etch is timed to produce a 20- μm gap, and is performed on only one of the two wafers in Figure 1. The remaining Si_3N_4 mask is then removed using hot phosphoric acid, and the wafer receives approximately 2 μm of sputtered gold. This gold thickness is slightly more than the 1.8- μm skin depth of gold at 2 GHz. Finally, the wafer is diced, and gold-gold thermocompression bonding of two die is used to form a closed cavity resonator (Figure 2C). Figure 3 shows a photograph of one half-cavity die before bonding; the die does not contain slits for the magnetic coupling loop.

KOH etching is important to cavity fabrication because it results in very smooth and open side walls, and open cavity corners. It is otherwise not possible to coat the cavity with a continuous low-resistivity film of gold. In contrast, deep-reactive-ion etching of silicon, and hydrofluoric-acid etching of glass, were not successful. The resulting cavity side walls were too steep and rough, and the resulting cavity corners were too sharp, to permit good metalization.

3. Modeling & Design

At 3 GHz, the wavelength of light is 10 cm. Since a 3-GHz resonator of the type shown in Figure 1 is much smaller than 10 cm, it is a quasistatic device [11]. Therefore, it has identifiable capacitive regions in which electric-field energy storage dominates, and inductive regions in which magnetic-field energy storage dominates. This argument holds for higher frequency resonators too because size will scale down as frequency scales up. In the resonator shown in Figure 1, the

Figure 2: cavity resonator fabrication process flow.

Figure 3: photograph of one half-cavity die.

central region functions as a parallel-plate capacitor while the surrounding toroidal cavity functions as a single-turn inductor. Together, the central capacitor and the toroidal inductor form a parallel resonant tank.

As fabricated in Figures 2 and 3, the capacitor has nearly square electrodes. Denote their side length by X , and the distance across their gap by G . Then, the capacitance of the tank, C_T , is approximated by

$$C_T \approx \frac{\epsilon_o X^2}{G} \quad (1)$$

where ϵ_o is the free-space permittivity. Fringing is ignored in (1) since $X \gg G$.

Also as fabricated in Figures 2 and 3, the inductor has a square outer perimeter. Denote its outer side length by W and the depth of its cavity by D . In this case, the mean path length of the solenoidal magnetic field through the toroid is approximately $(4X + \pi(W - X))/2$, and the cross sectional area through which the magnetic field passes is approximately $D(W - X + D \cot(54.7^\circ))/2$. The inductance of

the tank, L_T , is thus approximately

$$L_T \approx \frac{\mu_o D(W - X + D \cot(54.7^\circ))}{8X + \pi(W - X)} \quad (2)$$

where μ_o is the free-space permeability. Better approximations for L_T can be obtained from [12].

Since $D \ll (W - X)$ for the resonators considered here, the series resistance of the inductor is dominated by the resistance of the metal covering the floor and ceiling of the inductor cavity. In this case, the path length of the solenoidal current around the inductor toroid is approximately $(X - W)$, and the mean cross sectional area through which the current passes is approximately $\delta(X + W)/2$. The series resistance of the tank inductor, R_{LT} , is thus approximately

$$R_{LT} \approx \frac{W - X}{2\sigma\delta(W + X)} \quad (3)$$

where σ is the metal conductivity, and δ is the smaller of the metal thickness and its skin depth. Note that when δ is a skin depth, then R_{LT} is a function of frequency. Again, better approximations for R_{LT} can be obtained from [12].

To approximate the series resistance of the tank capacitor, model the capacitor electrodes as disks of radius R and thickness δ . Further, assume that the electrodes are uniformly charged, which is the case for a high- Q capacitor. In this case, the radial current density J in the electrode at the radius r from the center is $ri/2\pi\delta R^2$, where i is the total current leaving the capacitor. By integrating the power dissipation density J^2/σ over the volume of both capacitor electrodes, and equating the result to $R_{CT}i^2$, the series resistance of the tank capacitor, R_{CT} , is approximated by

$$R_{CT} \approx \frac{1}{4\pi\sigma\delta} \quad (4)$$

Note that when δ is a skin depth, then R_{CT} is also a function of frequency. Once again, better approximations for R_{CT} can be obtained from [12].

By combining (1) and (2), the resonant frequency of the tank, ω_T , is determined according to

$$\omega_T = \frac{1}{\sqrt{L_T C_T}} \approx \sqrt{\frac{G(8X + \pi(W - X))}{\epsilon_o \mu_o D X^2 (W - X + D \cot(54.7^\circ))}} \quad (5)$$

which is tuned by compressing the resonator to decrease G . Using (1) through (5) the quality factors of the capacitor, Q_{CT} , the inductor, Q_{LT} , and the tank, Q_T , are determined from

$$Q_{LT} = \frac{\omega_T L_T}{R_{LT}} \quad \& \quad Q_{CT} = \frac{1}{\omega_T R_{CT} C_T} \quad (6)$$

$$Q_T^{-1} = Q_{CT}^{-1} + Q_{LT}^{-1} \quad (7)$$

For the experimental resonator described below, $W = 18$ mm, $X = 4$ mm, $D = 1$ mm, and $G = 23.2 \mu\text{m}$, as fabricated for an uncompressed resonator. From (1), (2) and (5) it is therefore expected that $C_T = 6.11$ pF, $L_T = 243$ pH, and $\omega_T/2\pi = 4.13$ GHz. At this frequency, the skin depth of gold is $\delta = 1.22 \mu\text{m}$. From (3), (4), (6) and (7) it is therefore expected that $R_{LT} = 6.35$ m Ω , $R_{CT} = 1.59$ m Ω , $Q_{CT} = 3970$, $Q_{LT} = 920$, and $Q_T = 795$. In general, the inductor is always more lossy than the capacitor.

The wire loop and toroidal inductor form a transformer, and so the magnetically-coupled resonator is modeled as shown in Figure 4. There, α is the ratio of the loop cross sectional area to the inductor cross sectional area, $L_{L1} + L_{L2}$ is the leakage inductance of the loop, C_P models the distributed parasitic capacitance of the loop, and R_P models eddy currents induced by the loop in the inductor walls. From terminal impedance measurements alone, it is not possible to uniquely determine α , L_T , C_T , L_{P2} , and $R_T \equiv R_{LT} + R_{CT}$. Therefore, C_T will be taken to be 6.11 pF, as given by (1) for the as-fabricated dimensions.

Figure 4: electrical model of the experimental resonator.

4. Experimental Results

Resonator impedance is measured as a function of frequency with an Agilent PNA/E8362B network analyzer as shown in Figure 5. The resonator is mounted on a block, and a manipulator inserts the magnetic-coupling wire loop into the cavity as in Figure 1. The wire loop is kept away from the cavity walls to minimize the induction of parasitic eddy currents and losses in the walls. The wire loop connects to the analyzer through an SMA connector. Finally, impedance is measured as a calibrated piezoelectric actuator pushes on the resonator through a hole in the block to compress the cavity, thereby closing the capacitor gap and tuning the resonator.

Figures 6 and 7 show two impedance measurements, one with no cavity compression and one with a 13.9- μm compression of the capacitor gap. The figures show the tunable resonance of the tank, which shifts from 3.97 GHz to 2.54 GHz, as well as other dynamics caused by the wire loop. From

Figure 5: experiment to measure resonator impedance.

the measured impedance of the uncompressed resonator, and the assumption that $C_T = 6.11$ pF, the model parameters in Figure 4 are determined to be $L_T = 266$ pH, $\alpha = 0.431$, $L_{L1} = 769$ pH, $L_{L2} = 6.82$ nH, $C_P = 747$ fF, $R_T = 22.4$ m Ω and $R_P = 11$ k Ω . These parameters are very close to those predicted in Section 3. Further, the match between the modeled and measured impedance is extremely good up to and beyond 5 GHz, with less than a 1% error in magnitude and phase. Thus, the modeled and measured impedance would be indistinguishable in Figures 6 and 7, hence the modeled impedance is omitted. At frequencies above 6 GHz, the modeled and measured impedance diverges, most likely due to distributed wire loop dynamics that is not captured by the lumped-parameter model of Figure 4. Note that to compute the modeled resistances, the frequency-dependent skin depth of gold was used in (3) and (4).

Figure 6: resonator impedance magnitude as a function of frequency for 0- μm and 13.9- μm capacitor gap compression.

Figure 7: resonator impedance phase as a function of frequency for 0- μm and 13.9- μm capacitor gap compression.

With a capacitor gap compression of $13.9\text{ }\mu\text{m}$, (1) predicts a tuned C_T of 15.2 pF . With only this change to its parameters, the model of Figure 4 predicts the tuned impedance shown in Figure 6 with the same level of accuracy as it does the untuned impedance. Similar tests were repeated for all capacitor gap compressions from $1\text{ }\mu\text{m}$ to $13\text{ }\mu\text{m}$ in $1\text{ }\mu\text{m}$ steps with identical results. Thus, it can be concluded that the resonator behaves as predicted in Section 3.

Figure 8 shows the measured tank resonant frequencies that result from each $1\text{-}\mu\text{m}$ step in capacitor gap compression, together with a theoretical prediction based upon (1), (5) and the fitted value of L_T , 266 pH . The good match under-scores the validity of the analysis of Section 3. Figure 9 shows the corresponding measured Q of the resonator, and the estimated Q of the tank alone based on (7) and the fitted parameters of the model in Figure 4. Note that the Q of the tank is generally higher than that of the complete resonator owing to the losses induced by the wire loop. It is not, however, as high as predicted in Section 3 and Figure 4.

Figure 8: measured (*) and predicted (–) tank resonant frequency ω_T as a function of the decrease in capacitor gap.

5. Conclusions

The primary conclusion of this paper is that a cavity resonator can exhibit both high- Q , approaching 300, and a wide high-frequency tuning range. Further, it can be relatively simple to fabricate. In order to become acceptable to most applications, however, two modifications to the resonator as presented here must be made. First, direct electrical coupling to the capacitor should replace the magnetic coupling through a wire loop. The leakage inductance, and parasitic capacitance and resistance, of the wire loop appear to be generally too high. Second, the cavity structure must be made less rigid than that shown in Figures 1 and 3 so as to permit tuning with lesser mechanical effort.

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Figure 9: measured resonator (*) and estimated tank (□) quality factors at the resonance of ω_T as functions of the decrease in capacitor gap.

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