

DESIGN, SIMULATION AND TESTING OF PLANAR SPIRAL COILS FOR THE TIME-GATED INTERROGATION OF QUARTZ RESONATOR SENSORS

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ABSTRACT

A technique for contactless electromagnetic interrogation of quartz resonator sensors is proposed and validated. The technique is based on the separation in time of the excitation and detection phases, exploiting the sensing of the transient response of the resonator. Contactless operation is achieved by means of electromagnetic coupling between spiral planar coils connected to the interrogation circuit and the sensor. The typical operating frequencies of quartz resonators in the megahertz range limit the working range to short distances. Analytical modeling and numerical and finite element simulation of the geometry of the coils have allowed to extend working operating distances up to 10 cm.

INTRODUCTION

Quartz crystal resonators (QCR) are adopted as acoustic load sensors exploiting the piezoelectric properties of quartz crystals. Typical application is the employment as quartz crystal microbalance (QCM) for measurement in biochemical applications where the mass of substances to be detected can change the vibrating mass of the resonator and hence its resonance frequency (Janshoff and Steinem. 2001). More generally, a set of acoustic properties of QCRs can be affected by the load, resulting in complex changes in the resonance response (Benes et al 1995, Ferrari and Lucklum 2008).

The possibility to contactless interrogate the resonators via electromagnetic coupling is attractive for applications where cabled solutions are generally unpractical or not allowed, i. e. in enclosed environments like hermetic boxes or sealed food packages. One of the main issues in contactless measurement information is to grant the independence of the measurement on the distance between the interrogation unit and the sensor. In this perspective, adopting a resonant principle is a robust approach, being

the measurement information in the frequency of the vibrations of the resonator and hence virtually independent on the operating distance (Ogi et al. 2006, Baù et al. 2011). Either frequency-domain or time-domain approaches have been proposed for the contactless interrogation of QCRs. Frequency-domain techniques simultaneously excite and detect the resonators, measuring the impedance or a specific transfer function, but they can be affected by signal-to-noise ratio issues and in general the measurement is dependent on the distance between the measuring unit and the sensor. Time-domain techniques, as the one proposed in the present paper, typically separate in time the excitation and detection phases exploiting the transient response of the resonator. In particular, the frequency of the damped vibrations of the resonator is exploited, which in principle depends only on the mechanical parameters of the resonator. Contactless operation can be achieved adopting capacitive or electromagnetic principles, among others. Capacitive principles, though possible do not grant sufficient operating distance, due to the weakness of the electrostatic forces on operating distances of few centimeters. Contrary, an electromagnetic technique relying on the magnetic coupling between a primary coil used for both excitation and detection and a secondary coil connected to the resonator (Lucklum and Jakoby 2009, Wu et al. 2008) has been adopted. The operating frequencies of QCRs are typically in the range of tens of megahertz and, as a consequence, the electromagnetic coupling operates on short-range distances between the primary and secondary coils.

To obtain suitable interrogation distances at feasible size of the primary and secondary coils, i.e. ultimately of the sensor element, it is of critical importance the design of the coils and simulations have a prominent role in this task. Downscaling of the dimensions of the sensor could also be beneficial for sensing applications where unobtrusiveness may be of first concern. Additionally, the measuring technique proposed herein does not need on-board power supplies, because the sensor operation relies on the intrinsic piezoelectric properties of quartz crystals and on electromagnetic coupling. In this perspective, the adoption of batteries

and related issues of maintenance and/or periodic replacement can be advantageously avoided. This paper is devoted to theoretical investigations and numerical simulations of the geometry of planar coils and their coupling. The designed coils have been realized and the simulation results have been verified by experimental tests.

OPERATING PRINCIPLE

The operating principle of the proposed interrogation technique is depicted in Figure 1. It relies on the separation in time of the excitation and detection phases. The system is composed of a primary coil L_1 which can be alternately switched between a signal generator and the input of a measuring amplifier. The primary, or TX, coil L_1 is magnetically coupled to a secondary, or RX, coil L_2 which is connected to the QCR sensor. During the excitation phase the switch is in position E and the sinusoidal signal $v_e(t)$ at frequency f_e is applied to the coil L_1 . The current in L_1 couples a magnetic field on coil L_2 and hence a current i_2 circulates in the QCR sensor connected to L_2 . The current i_2 excites the QCR into vibrations at frequency f_e . The frequency f_e is chosen to be close to the thickness-shear first resonant mode f_m of the QCR sensor, even if this is not strictly required because the proposed principle relies on the detection of the free decaying response of the resonator which is independent of the excitation frequency. This is advantageous because the resonant frequency of the QCR cannot be known in advance. The ideal condition $f_e=f_m$ enhances the signal-to-noise ratio during the detection phase. When the switch is set to position D after a period of time T_E , the QCR sensor undergoes decaying harmonic oscillation to its damped resonant frequency f_{dm} . This oscillation, thanks to the piezoelectric properties of quartz, induces a current into L_2 which in turn couples to L_1 . The induced voltage $v_o(t)$ on L_1 is read by means of an amplifier. The signal $v_o(t)$ is fed to a zero-crossing detector to derive a square wave signal at frequency $f_o=f_{dm}$ measurable with a frequency counter. If the

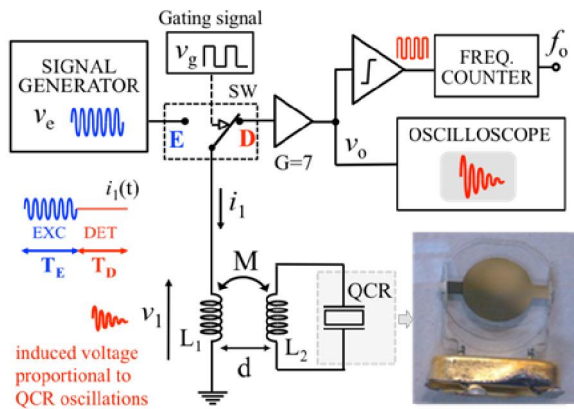


Figure 1: Schematic diagram of the operating principle.

mechanical quality factor Q of the QCR sensor is high, then $f_o=f_m$. The readout signal is sensed for the period T_D after which a new excitation phase starts. Considering high- Q QCR sensors, the voltage $v_1(t)$ sensed across the inductor L_1 during the detection phase can be approximated by (Bau et al. 2011):

$$v_1(t) = MA_m \exp(-t/\tau_m) \cos(2\pi f_{dm}t + \theta_m) \quad (1)$$

where M is the mutual inductance between the coil L_1 and L_2 , A_m and θ_m are the amplitude and phase terms which depend on the initial mechanical and electrical conditions at the beginning of the detection phase. The exponential decaying time τ_m is related to the mechanical quality factor by $Q=\pi f_m \tau_m$ while the damped resonance frequency is related to the mechanical resonance frequency by $f_{dm}=(4\pi^2 f_m^2 - 1/\tau_m^2)^{1/2}$. In (1) it can be observed that the mutual inductance M acts as a scaling factor, without affecting the frequency of the readout signal. On the other side, the value of M depends on the distance, orientation and geometry of the coils and therefore sets the maximum operating distance of the proposed interrogation principle. Hence, the design of the proper geometry of the coils is fundamental to achieve larger operative distances and it is subject to dimensional, electrical and mechanical constraints.

INDUCTIVE COUPLING

To optimize the operating distances between the primary and secondary coils, the influence of the geometrical and electrical parameters of the coils on the electromagnetic coupling have been investigated by means of theoretical methods as well as by numerical simulation.

Coil equivalent circuit

Several closed-form expressions have been proposed to approximate the inductance L of printed spiral coils (PSCs). We adopted (2) from (Mohan 1999) for square-shaped coils whose layout is shown in Figure 2 on the left

$$L = \frac{2\mu_0 n^2 l}{\pi} \left[\ln\left(\frac{2.067}{\phi}\right) + 1.078\phi + 0.125\phi^2 + 0.5 \frac{(n-1)s^2}{(\phi)^2} \right] + \frac{2\mu_0 n^2 l}{\pi} \left[0.178 \frac{(n-1)s}{nl} + 0.0833 \frac{(n-1)s(s+w)}{l^2} - \frac{1}{n} \ln\left(\frac{w+t}{w}\right) \right] \quad (2)$$

$$\phi = \frac{d_{out} - d_{in}}{d_{out} + d_{in}} \quad (3)$$

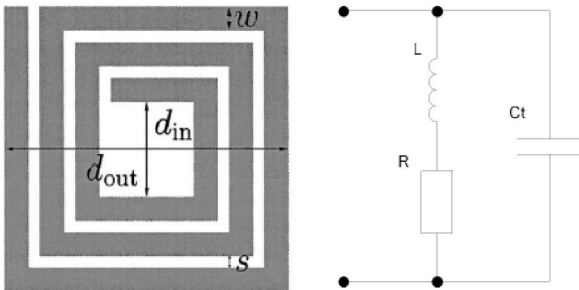
where n is the number of turns d_{out} and d_{in} are the outer and the inner diameters of the coil, μ_0 is the vacuum permeability of space. The wire cross-section is rectangular and its width and thickness are w and t , respectively. The wire sections are uniformly spaced with edge-to-edge spacing s ; φ is a parameter known as fill factor, which changes from 0, when all the turns are concentrated on the perimeter like filament coils, to 1, when the turns spiral all the way to the center of the coil.

The right panel of Figure 2 shows the equivalent resonant circuit of the PSC. The series resistance is calculated using the following expression

$$R = \frac{\rho l}{w\delta(1 - e^{-\frac{t}{\delta}})} \quad (4)$$

$$l = 4nd_{out} - 4nw - (2n + 1)^2(s + w) \quad (5)$$

where l is the length of the conductive trace, ρ is the resistivity of the conductive material and δ is the metal skin depth. It must be emphasized that the AC-resistance is also affected by proximity effects caused by the currents that are flowing through one or more other nearby conductors, such as within a closely wound coil of wire. The total capacitance is affected by the presence of two insulating materials: one is the air filling the gap between adjacent traces and the other is the PSC substrate, which could be ceramic, polyimide, or FR4. In literature there are many analytical approximations for the parasitic capacitance, but the estimation accuracy can be insufficient, therefore it becomes necessary to use the available simulation tools to validate the design result. In this study, 3-D numerical simulations were used for an accurate prediction of the parasitic elements and for the final tuning of the proposed PSCs. Even though many fabrication processes allow using multilayer conductors, we limited our design to single layer PSCs because the parasitic capacitance for multilayer PSCs is much larger than for single layer PSCs and this can significantly reduce the self-resonance frequency (SRF).



Figures2: Geometrical parameters of a square-shaped printed spiral coil. Schematic drawing (on the left) and equivalent resonant circuit (on the right).

Coils mutual inductance

The derivation of the general mutual inductance formula for planar structures starts from the mutual inductance between two circular air cored loops whose axes are parallel; by using Maxwell's equations, one can obtain

$$\begin{aligned} M(a, b, \gamma, d) &= \pi\mu_0\sqrt{ab} \int_0^\infty J_1\left(x\sqrt{\frac{a}{b}}\right)J_1\left(x\sqrt{\frac{b}{a}}\right) \\ &\times J_0\left(x\frac{\gamma}{\sqrt{ab}}\right)\exp\left(-x\frac{d}{\sqrt{ab}}\right)dx \end{aligned} \quad (6)$$

where a and b are the radii of the two coils, d is the separation between the coils and γ is distance between their axes. J_0 and J_1 are the Bessel functions of the zeroth and first order.

This expression does not contain the radius R of the coil's wire since it is assumed that the ratios R/a and R/b are sufficiently small (Zierhofer and Hochmair 1996). For perfectly aligned coaxial coils, where $\gamma = 0$, (6) can be simplified to the Neumann's expression treated in numerous text books (Jow and Ghovanloo 2007):

$$M(a, b, d) = \frac{2\mu_0}{\alpha}\sqrt{ab}\left[\left(1 - \frac{\alpha^2}{2}\right)K(\alpha) - E(\alpha)\right] \quad (7)$$

where $K(\alpha)$ and $E(\alpha)$ are complete elliptic integrals of the first and second kind, respectively and where

$$\alpha = 2\sqrt{\frac{ab}{(a+b)^2 + d^2}} \quad (8)$$

Since a PSC can be considered as a set of concentric single turn coils with shrinking diameters and connected in series, we can use (5) or (6) to calculate the overall M of two PSC in parallel planes: the mutual inductance value can be found by summing the values of the partial mutual inductances between every turn of one coil and all the turns of the other coil

$$M_{tot} = g \sum_{i=1}^{n_1} \sum_{j=1}^{n_2} M_{ij}(a_i, b_i, \gamma, d) \quad (9)$$

where g is a factor dependent on the shape of the PSC and in (Jow and Ghovanloo 2007) they found empirically that $g \approx 1.1$ for a square-shaped PSCs.

In this study, the primary coil (TX) and the secondary coil (RX) are separated by a distance d which is typically in the range of a few centimeters. The coils are loosely coupled and are characterized by extremely low coupling coefficients ($k \leq 0.01$) (Fotopoulou and Brian 2011). As a matter of fact, this happens when the desired separation (d) between the TX and RX coils is larger than the coil dimensions as is the case in most of radio frequency identification (RFID) applications.

For this separation values, we can take advantage of the properties of the magnetic field at large distance and we can assume a homogeneous magnetic field. Under this

hypothesis, the mutual inductance M between two distant square spiral coils in coaxial configuration can be readily calculated:

$$M = \frac{B(0,0,d)A_{RX}}{I} \quad (10)$$

where A_{RX} is the area of the RX coil and I is the current that excites the TX coil. The dominant B_z magnetic field component at the center of the RX coil can be obtained by integrating Biot-Savart's law around the TX coil having side length L

$$\vec{B}(0,0,z) = \frac{\mu_0 I}{2\pi \left(\frac{z^2}{L^2} + \frac{1}{4}\right) \sqrt{z^2 + \frac{L^2}{2}}} \hat{z} \quad (11)$$

When the RX coil is distant from the reader antenna (further than a side length L , such that $z \gg L$), the magnetic field falls off rapidly:

$$\vec{B}(0,0,z) \approx \frac{IL^2}{2\pi z^3} \hat{z} \quad (12)$$

The following expressions can be generalized for a PSC with n turns where $L=d_{in}$ and from (12) we can conclude that the optimal value for d_{in} depends on the coils relative distance, d .

We demonstrated by comparing (10) and (7) for a set of PSCs varying the distance of separation that $d > d_{out}$ is enough for considering a homogeneous magnetic field at the RX coil.

We have used equations (2)-(11) in combination with CST (Computer Simulation Technology) and COMSOL Multiphysics to find the optimal coil geometries. We have also fabricated a number of PSCs on FR4 substrate, designed through a procedure which will be explained in the next section.

Simulation of coil geometries

The self-inductance of the TX and RX coils required by the sensor circuit at the operation frequency of 4.8 MHz is 8 μ H and the desired reading distance is more than 5 cm. In order to design the coil and optimize the interrogation range we developed a MATLAB code devised for coils made of 1-D filaments wires by sweeping the parameters included in (2), (4) and (10). Furthermore, we tried to maximize the mutual inductance M by designing a TX PSC with d_{in} as large as possible according to (11). We fixed s (200 μ m), and t (35 μ m) at the minimum value permitted by the fabrication technology even if this choice may not be optimal. In (2) and (4) we swept n and d_{in} for a fixed inductance value L (8 μ H) and we selected the TX PSC ($n=6$) with a high d_{in} (69.2 mm) which we analyzed and optimized numerically with CST. Note that at high frequencies (>10 MHz) the effects of PSC parasitic capacitance becomes more significant for PSCs with larger diameters and the self-resonance frequency (SRF) moves downward quickly; w is 400 μ m to guarantee low

resistance loss ($\approx 3\Omega$) within a large frequency range including 4.8 MHz.

After selecting the geometrical parameters of the TX PSC we focus on the secondary PSC. Ideally, for a compact reading device, the d_{out} RX PSC should be comparable to the sensing passive device. The sensor transmits in the detection phase so we have to proceed as before to reach a large operation distance. Table I shows the geometries and specifications of four square-shaped PSCs that were designed and then fabricated using 35 μ m thick copper on FR4 substrates. PSC2~4 were used as RX PSC and PSC1 as RX-TX PSC. Figure 3 shows the trend of the mutual inductance versus the distance of separation for fabricated coils in Table I calculated numerically by the simulations performed in COMSOL Multiphysics. The optimization of the mutual coupling considered not only the aligned coaxial coils but also the presence of misalignment between the coil axes. Figure 4 shows the calculated mutual inductance values for a distance $d=6$ cm as a function of the displacements in the (x,y) plane perpendicular to the coil axis (along z).

EXPERIMENTAL VALIDATION

Planar coils for the experimental test have been fabricated by milling a FR4 substrate with 35 μ m-thick

Table I: Specifications of the PSCs used in measurements

Parameter	PSC1	PSC2	PSC3	PSC4
Shape	Square	Square	Square	Square
d_{out} (mm)	76	34.3	31.2	30
d_{in} (mm)	69.2	17.9	10.1	1.6
n(turns)	6	14	18	24
w(μ m)	400	400	400	400
s(μ m)	200	200	200	200
L(μ H)	8.027	8	8.35	8.5
R(Ω)	3	2.52	3.1	2.65
SRF(MHz)	22.4	48	50	54

copper layer. The geometric properties of the coils are the same reported in Table 1. In particular, two coils corresponding to PSC1 in Table 1 have been considered. A picture of the coil is shown in the inset of Figure 5. By means of a HP4194A impedance analyzer the mutual inductance M between the two coils at various distances has been measured. Figure 5 compares the values of the mutual inductance measured at 6 MHz with the values predicted by COMSOL simulations. Additionally, the system described in Fig. 1 has been tested adopting a QCR with resonance frequency of about 4.8 MHz as the sensor. Multiple coils have been fabricated according to the geometrical parameters of Table 1. In particular, a PSC1 coil has been connected to the interrogation circuit during all the experimental tests, while the sensor has been connected alternatively to all the PSC1~PSC4 coils.

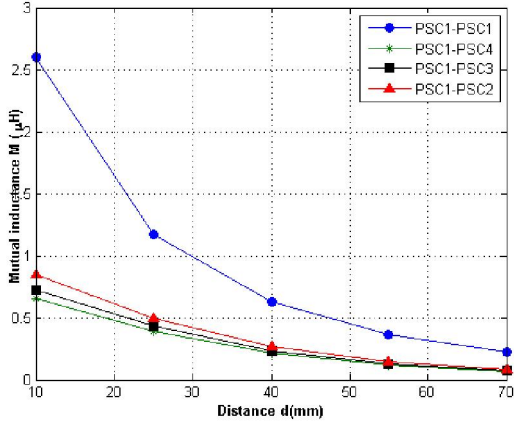


Figure 3: Mutual inductance versus the separation distance d for the fabricated coils of Table I (values calculated by means of COMSOL Multiphysics).

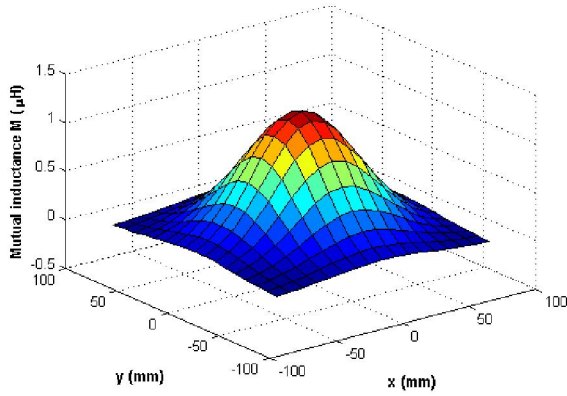


Figure 4: Mutual inductance computed for PSC1-PSC2; PSC1 is centered at the origin and is parallel to the (x,y) plane and PSC2 moves in the (x,y) plane at the separation distance $d=6$ cm.

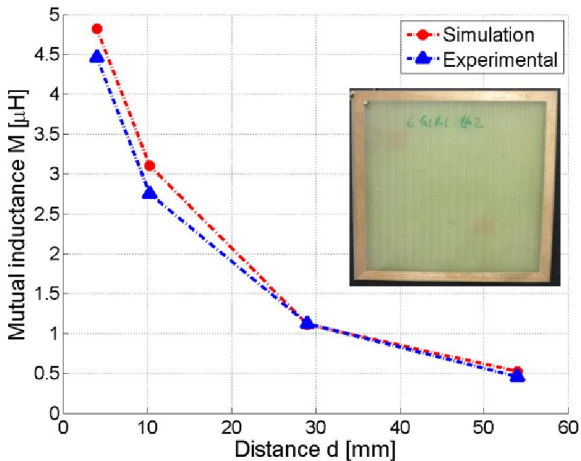


Figure 5: Comparison between the simulated and measured values of the mutual inductance M for the PSC1 coil.

The transient response of the sensor has been acquired by means of a digital oscilloscope Agilent

MSO-X3014A starting from a prescribed delay after the gating signal has switched to the detection phase. The acquired signal FFT has been calculated and the amplitude of the peak in correspondence of the sensor resonance frequency has been plotted as a function of the interrogation distance d , as shown in Figure 6. As it can be observed, the trend of the experimental points for each of the coil pair follows the trend of the simulated values of the mutual inductance M of Figure 3, confirming that M acts as a scaling factor on the response of the sensor.

CONCLUSIONS

A technique for the contactless electromagnetic interrogation of QCR sensors which alternatively excites the sensor and detects the transient response, has been proposed. The combination of the time-gated interrogation technique with the adoption of a resonant sensing principle is robust with respect to the distance between the interrogation and sensor unit because the measurement information is carried by the resonance frequency of the resonator. The typical operating frequencies of QCRs are in the megahertz range, confining the operation of the proposed technique to short-range distances. To enhance performance in terms of operating distances, we optimized the geometries of printed spiral coils adopting a semi-empirical method, using a combination of theoretical results (1-D filament) and 3-D numerical simulations.

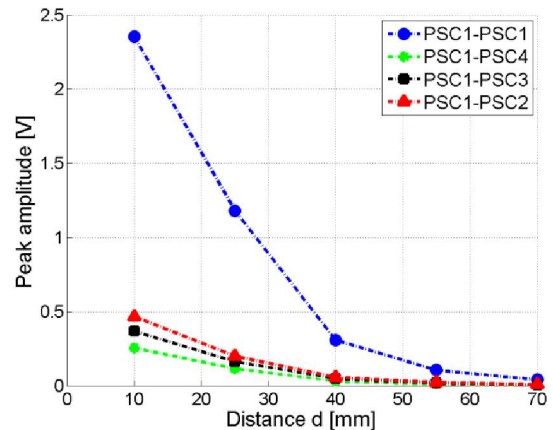


Figure 6: Amplitude of the FFT magnitude peak in correspondence of the resonance frequency, measured by means of the interrogation circuit for a subset of the coil pairs of Table I

This method takes into account the effects of the parasitic elements (resistance and capacitance) on the operative band and mutual inductance of the designed coil. The obtained experimental results confirm the predictions of the lumped element equivalent model in Figure 1, demonstrate the effectiveness of the approach and show an optimized working distance of up to 10 cm in the case PSC1-PSC1. The proposed system can be exploited for the measurement of physical/chemical quantities affecting the resonant response of QCR sensors.

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