

PARTICLE SENSING WITH NARROWBAND MICROWAVE SPLIT-RING RESONATORS

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By
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PARTICLE SENSING WITH NARROWBAND MICROWAVE
SPLIT-RING RESONATORS

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September 2021

We certify that we have read this thesis and that in our opinion it is fully adequate,
in scope and in quality, as a thesis for the degree of Master of Science.

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ABSTRACT

PARTICLE SENSING WITH NARROWBAND MICROWAVE SPLIT-RING RESONATORS

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A new microwave sensor with concentric split-ring resonator (CSRR) topology is designed and integrated with microfluidics. While the resonator is being driven at its resonant frequency, cells passing through the channel shift this frequency. Information regarding the cell (or microparticle) can be extracted from the amplitude of this shift. The design is based strongly on simulations and shows good agreement with the theory. Experiments are performed to evaluate the sensor's performance with polystyrene particles. These experiments show that the sensor gives an outstanding SNR value: around 140kHz mean signal amplitude compared to a 1.2kHz average noise amplitude. To further evaluate the performance, single-cell experiments are performed. A target cell is selected and passed through the sensing region repeatedly. Frequency shifts are recorded. The mean frequency shift is 44kHz, and the signal-to-noise ratio is over 140. The ensemble standard deviation for the frequency shifts is 9.33kHz, and the variance in the results can be further improved with creative microfluidic designs.

Work has been directed towards improving the measurement setup for split-ring resonators. A split-ring resonator sensor is designed and integrated into an oscillator loop, designed from scratch. The design criteria for this project are; faster acquisition times, control-loop parameter independence, and improved cost-effectiveness. It is shown that the new oscillator setup satisfies all the design criteria. The design process is once again simulation-driven. Real-life experiments are also performed with the oscillator boards for performance evaluation. Allan deviation experiments show promising results (1.49×10^{-6}) regarding oscillator stability. Polystyrene experiments show a lower response compared to the CSRR sensors.

Keywords: Ring resonators, Microwave, Biophysical sensors, Single cell detection.

ÖZET

DAR-BANTLI MİKRODALGA BÖLÜNMÜŞ HALKA REZANATÖRLERİ İLE PARTİKÜL ALGILAMA

Berk Küçükoglu

Makine Mühendisliği, Yüksek Lisans

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Mikrodalga rezonatörler, herhangi bir etiketleme olmaksızın tek hücreleri gerçek zamanlı olarak algılamak için güçlü bir platform sağlar. İlk bölümde, iç içe geçirilmiş bölünmüş halka rezonatörü geometrisine sahip yenilikçi bir mikrodalga sensörü tasarlanmıştır. Aynı zamanda mikro akışkan kanallar ile entegrasyonu sağlanmıştır. Rezonatör rezonans frekansında sürülürken, kanaldan geçen hücreler bu rezonans frekansı değerini kaydırır. Burada frekans kayması genliğine bakılarak, geçen hücre (veya mikroparçacık) ile ilgili bilgi edinilebilir. Bu sensörün tasarım aşamasında çoğunlukla simülasyonlar kullanılmıştır. Bu şekilde yapılan sensörden alınan sonuçlar, teori ile de uyum göstermektedir. Sensörün performansını değerlendirmek amacıyla polistiren partiküller ile deneyler yapılmıştır. Polistiren parçacıkların kanaldan geçişi ortalama 140kHz kayma sağlarken, gürültü seviyesi ortalama 1.2kHz değerindedir. Performansı gerçek koşullarda değerlendirmek amacıyla, tekil hücre deneyleri de yapılmıştır. Bu deneylerde bir tane hücre seçilmiştir ve aynı hücre tekrar tekrar hassas bölgeden geçirilmiştir. Bunun sonucunda olan frekans kaymaları kaydedilmiştir. Ortalama frekans kayması 44kHz, sinyal ve gürültü oranı ise 140 değerinin üzerindedir. Verilerdeki kümeye ait standart sapma ise 9.33kHz değerindedir. Yaratıcı mikroakışkan kanalı tasarımları ile bu varyasyonlar azaltılabilir.

İkinci bölümde ise, ölçüm kurulumunu iyileştirmeye yönelik çalışmalar yapılmıştır. Bu sefer bölünmüş halka rezonatörü geometrisine sahip bir sensör üretilmiş ve transistör seviyesinden kurulan bir osilatör ile entegrasyonu sağlanmıştır. Bu konfigürasyon sayesinde, tasarım basamağında belirtilen tüm gerekler sağlanmıştır. Örnek olarak, osilatör entegrasyonlu sensörden bilgi almak çok daha hızlıdır. Aynı zamanda sensörün başarısının parametre ayarına

olan bağımlılığı da ortadan kaldırılmıştır. Tasarım süreci yine simülasyonlara dayalıdır. Performans değerlendirilmesi için Allan sapması deneyleri yapılmıştır. Bu deneyler sonucunda umut verici (1.49×10^{-6}) Allan Sapması sonuçları elde edilmiştir. Polistiren parçacıklar ile yapılan deneylerde ise ilk çalışmada olan başarılar yakalanamamıştır.



Anahtar sözcükler: Halka rezonatörler, Mikrodalga, Bio-fiziksel sensörler, Tekil hücre görüntüleme.

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Chapter 1

Introduction

This thesis presents two sensors based on split-ring resonators and evaluates their performance with real-life experiments. First, a concentric split-ring resonator (CSRR) sensor is introduced. Then an oscillator integrated split-ring resonator sensor is demonstrated. Both chapters include detailed design processes, simulations, and experiments.

1.1 Detection with Concentric Split Ring Resonators

Optical visualization techniques are very well established and used through the science community to understand biological cell dynamics. For example, optical microscopy is used [3] for understanding cell growth. In another work [4], cell mass is observed through quantitative phase microscopy techniques. However, these methods are based on the “optical image” of the cell. Information regarding the weight, density, electrical permittivity, and other physical characteristics are “estimated” through these images. With the recent advances in biology, the need for sensors capable of measuring these less-accessible quantities has emerged.

Among these required quantities, mass and volume can be very informative metrics to evaluate and understand cell-related attributes. Micromechanical and nanomechanical structures have been used for many years to perform measurements at single-cell mass levels. A noise floor of approximately 20 zeptograms in vacuum has already been achieved [5] as early as 2006. Hence the challenge is not solely about mass sensitivity. The problem lies in integrating two systems: the high-precision mass measurement setup and the microfluidic setup responsible for guiding the cells in their natural (wet) environment. There are various works [6, 7] which bring the two together, yet the sensitivity is lost due to viscous damping, and throughput is low due to inadequate diffusion rates. To overcome these problems, creative solutions have been implemented: Manalis group developed [8, 9, 10] and improved [11, 12, 13, 14, 15] the suspended microchannel technology. Here, the mechanical cantilever does not resonate in the fluid environment. Instead, the cells and nanoparticles (the fluid) travel inside the cantilever. Although this is an exciting innovation, there are major challenges in the fabrication process of a hollow cantilever and the associated costs.

A less commonly investigated physical quantity is the electrical volume of the cell. Even though this is not a standard physical quantity, it can be very informative in many scenarios. Electrical volume is a function of dielectric permittivity. The electrical permittivity difference between the cell and its medium is due to proteins and biochemical material found inside the cell. Mass measurement techniques treat the cell as a whole without much information regarding what is contained inside. From the perspective of a mass sensor, a polystyrene particle and a cell weighting the same are identical entities.

Electrical sensors, with necessary conditions satisfied, can extract information regarding the material contained within single cells. The condition is mainly related to the frequency of operation. Most electrical sensors measure the decrease in complex conductivity induced by a cell or microparticle passing through a channel. Examples of these sensors are Coulter Counter [16] and impedance spectrometer [17] type of devices. At the operation frequencies of these sensors (typically < 30 MHz), the sensing process is dominated by polarization-related effects, known as the Debye screening. EM waves cannot penetrate into the cell,

and the cell is visualized (in an electrical signal sense) as a sphere-like opaque particle.

In this chapter of the thesis, a microwave sensor in the form of a Concentric Split-Ring Resonator is proposed to create a sensor that can probe inside of cells. To make this possible, operation frequencies have been adjusted to the GHz range, where the cell membrane becomes transparent [18]. Compared to earlier Bowtie design [19] performing a similar task, the CSSR sensor gives at least ten-folds better performance. With these improvements, the aim of achieving low-cost, high-sensitivity Microfluidics-Integrated Microwave Sensors (MIMS) is becoming much more realistic.

1.2 Micro-particle Detection with Standalone SRR Integrated Oscillators

The future of personalized medicine lies in the access to low-cost, high-throughput, user-friendly platforms that can detect and characterize (both in and out) cells in real-time. There is currently a vacancy for a system that can satisfy these three criteria at once. As an example, there are very sensitive methods [20] to characterize cells in real-time. However, they require laboratory-grade equipment and trained personal to operate. Cost-effectiveness is another problem. Although NEMS and MEMS devices have excellent sensitivity, they usually require a vacuum to achieve their top performance. The requirement of a vacuum makes the method expensive and less accessible. The ability to probe inside the cell is also desirable, as explained in the previous section.

The CSRR sensor was capable of satisfying many of these criteria, except being low-cost and user-friendly. It required expensive devices to operate, such as a lock-in amplifier. Furthermore, the controller parameters were required to be set by an expert before each experiment. Response due to cell passage events was strongly dependent on these parameters.

In this chapter, a sensor similar to CSRR is designed and integrated with an oscillator setup. This time, the Split-Ring Resonator geometry is used. The sensor is no longer driven by an external source, as it is integrated with a single-transistor low noise amplifier to form an oscillator loop. The cost is much lower than the CSRR setup, as it does not require lock-in amplifiers or complex up-conversion circuitry. The throughput is enhanced as the delays due to communication between instruments are eliminated. Lastly, the sensor's response is no longer parameter-dependent, considering there are no control loops in the system.

Chapter 2

Detection with Concentric Split Ring Resonators

Electrical sensors, such as Coulter counters and impedance spectrometry, used in lab-on-a-chip applications generally measure the decrease in complex conductivity induced by a cell or microparticle passing through a channel. At the operation frequencies of these sensors (typically < 30 MHz), Debye screening limits the penetration of electrical signals into the analytes. As such, the signal measured with these sensors, to first order, depends on the volume of the particle (which blocks the conductive channel) but not their internal material structure. However, probing the material properties, such as dielectric permittivity, can open many applications currently inaccessible with electrical sensors.

Microwave resonators provide a powerful platform to detect single cells in real-time without any form of labeling. SRRs (split-ring resonators) have become popular in recent years due to their lower radiation loss. To increase the sensitivity, such resonators can be enhanced by fabricating sensing regions where the gap between the two electrodes of the resonator narrows down. Cells passing through these regions have a significant impact on the resonant frequency. These shifts in the resonant frequency contain information about the cell's volume and internal dielectric properties. As a result, microwave sensors can enable the measurement

of permittivity differences of particles and differentiate between particles of the same size composed of different materials.

2.1 Design and Theory of Operation

The main idea is to develop a high-frequency EM sensor with a cell-sized discontinuity. A gap of this size has two benefits: First, it focuses the electric field to a tiny volume. Secondly, as the gap is cellular-sized, cells passing through this split will have a massive effect on the property of the medium in between. The passage of cells is similar to replacing all the material between the plates of a parallel-plate capacitor. For the analogy applied to our case, the removed dielectric is water, replaced by the biological material found on and inside the cell. The passage affects the resonant frequency of the EM resonator, and as the resonant frequency is tracked through time, information regarding the characteristics of the passing cell is gathered.

Initially, the design of the Concentric Split-Ring Resonator (CSRR) was based on microstrip ring resonators. Before moving onto ring resonators, here is a brief review of the microstrip circuits based on the book [1] “Foundations of Microstrip Circuit Design.” In-depth information on the following subsection can be found there.

2.1.1 Preliminary on Microstrip Circuits and Series Gap Discontinuity

Microstrip circuits consist of a conductor signal layer on top, a conductor ground layer at the bottom, and a dielectric substrate in between these layers.

Although having a simple-looking geometry, the electromagnetic fields associated are somewhat complicated. This complication is due to the sudden change

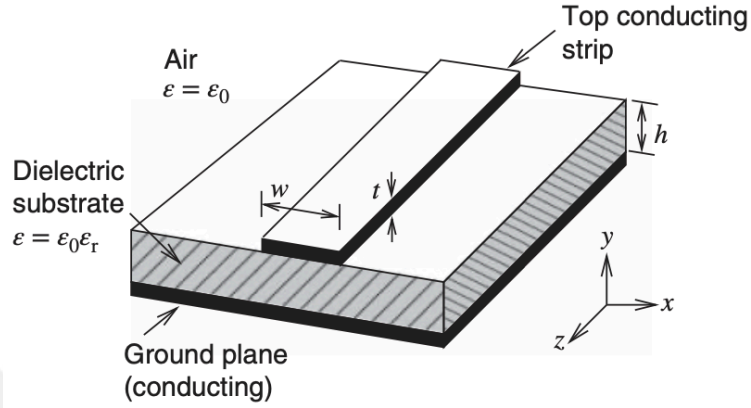


Figure 2.1: Illustration of microstrip circuit topology (taken from [1]).

of dielectric material from the substrate to the air. Due to this, microstrip transmission lines cannot support a single mode of propagation [21]. Although this is true, the majority of the transferred energy is tolerably close to being transverse. Hence, the field distributions in a microstrip transmission line are called quasi-TEM.

Microstrips are the most commonly used transmission line geometry in microwave engineering. Apart from their role as simple transmission lines for energy transfer, they can also be designed to imitate passive lumped elements like capacitors and inductors. These types of circuits are called distributed circuits.

For our case, we will focus on the series gap discontinuity. Although there are other approaches to model the series gap, it can be modeled with three capacitors [22] especially at frequencies below 10 GHz. These capacitor models are intuitive concerning the way electrical field lines behave.

Assuming the width of the line is equal for both sides of the discontinuity, the capacitors named C1 represent the fields fringing down to the ground layer. The capacitor C2 is the capacitance across the gap, formed similarly to a parallel plate capacitor.

The series gap discontinuity is vital for some forms of filters, coupling to resonators and other circuits. In our case, this discussion will be meaningful during

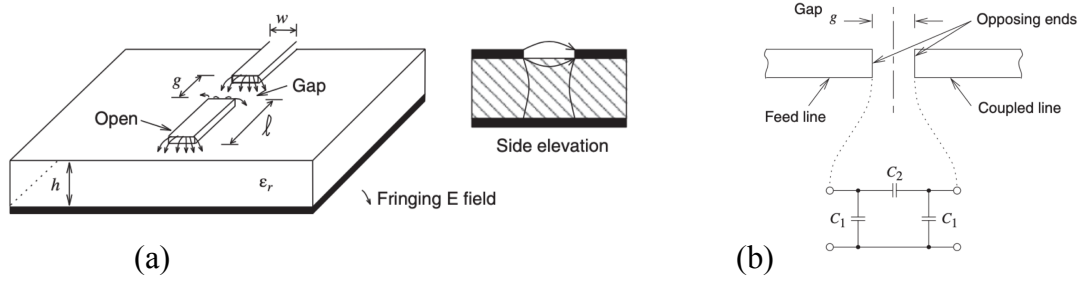


Figure 2.2: (a) Series gap with electric field illustration and (b) corresponding model schematic. Both figures taken from source [1].

the design process, which comes in the next section.

2.1.2 Constraints for the Design Process

As in every physical design process, there are not many dimensions that are settled from the beginning. There are only some boundaries loosely guiding the design process. With ring resonators, there are three parameters to define:

- Diameter of the resonator ring
- Width of the resonator trace
- Coupling gap to the drive and read circuits

Appropriate parameters must be chosen for a successful design. The parameters are chosen with respect to these four main ideas:

1. The resonator section, which is responsible for detecting cells, should be isolated from outside as much as possible. Coupling between the resonator and circuitry responsible for information extraction should be low, keeping the loading effects to a minimum.
2. The resonance frequency should be suitable for cell detection. It should not be too high that cells become entirely transparent. It should also not be

too low that the cell wall becomes opaque due to Debye screening, which will be explained below.

3. Most of the electric field should be confined to a small region where the sensing will occur. Any field that radiates away is a loss.
4. For low-noise operation, the resonant frequency should be away from common electromagnetic emission sources, such as Wi-Fi at 2.4 GHz and similar.

In the following section, these problems are tackled using an iterative approach.

2.1.3 Guided Design Process: Ring Resonator Base

The sensor should be free to resonate without any loading effect. With some hypothetical method, energy should be pumped into the resonator, and information regarding the resonant frequency should be extracted without loading the resonator at all. This way, cells passing over the resonator will have the full effect of changing the resonant frequency. As for a counter case, if the feedline to excite the sensor is directly connected to the resonator through a solid transmission line (shorted), the resonator will be over-coupled to the driving circuitry, and the resonant peak will fade away. In this case, goal of tracking the resonant frequency may not be reached.

However, the "completely unloaded resonator" idea is an unrealistic approach. In order to drive the resonator and extract information, there should always be some form of coupling in between. After other parameters are decided upon, the coupling distance will be swept to find an optimal value based on the quality of the resonant peak.

Since the sensor will have the ring resonator geometry, the formulas for calculating the resonant frequency are well-known [2]. With the most straightforward approach, the resonant frequency of a ring resonator in the absence of coupling

effects can be calculated using the formula,

$$2\pi r = n\lambda_g, \text{ for } n = 1, 2, \dots, \quad (2.1)$$

where r is the radius of ring resonator, and λ_g is the guided wavelength. λ_g is both frequency and material dependent. Equation to calculate the guided wavelength is

$$\lambda_g = \frac{1}{f\sqrt{\mu\varepsilon}}, \quad (2.2)$$

where μ and ε are permeability and permittivity, respectively. Our material of choice is SiO_2 , which we will have a discussion about. From the material database [1], the dielectric permittivity is found as:

$$\varepsilon = 3.8 (\varepsilon_0) = 3.8 (8.854 \times 10^{-12}) [\text{F/m}]. \quad (2.3)$$

The relative permeability value for most dielectrics is unity. Using these values, an initial design can be established. Inserting equation 2.1 into 2.2, we get a design formula:

$$r = \frac{n}{2\pi f\sqrt{\mu\varepsilon}}. \quad (2.4)$$

Given the desired frequency of operation, the resulting radius of the ring resonator can be calculated.

Now we can choose a suitable resonant frequency. For this, there are three (partially conflicting) constraints. The operation frequency should be:

- (i) Suitable for cell detection
- (ii) High enough to have a compact sensor
- (iii) Away from common sources of radiation

First and foremost, the proposed sensor design operates based on the dielectric permittivity contrast between external medium and biochemical components. We want our sensor to probe inside the cell, meaning the electric fields have to penetrate through the cell membrane. At low frequencies, detection is limited by

Debye screening. As an electric field is applied, ions in the membrane become polarized. Hence for low-frequency electric fields, the cell membrane is opaque [18]. On the other hand, the dielectric permittivity difference between biochemical components and the surrounding medium lessens at very high frequencies, making detection of the bio-components challenging.

In addition to this, operation at low frequencies requires extremely large resonators. For a rugged design, the frequencies should be kept high. For example, the guided wavelength is 77 mm at 2 GHz, 154 mm at 1 GHz, and 308 mm at 500 MHz. Having a ring resonator with a fundamental resonant frequency of 500 MHz requires a diameter of approximately 100 mm, which is unreasonable for microwave resonators. The reason behind this will also become evident in the manufacturing section.

With all of the constraints in place, a suitable suggestion is a (fundamental) operation frequency of 2.7 GHz. Using the design formula, the corresponding ring resonator should have an average diameter of 18mm.

	250μm	500μm	1000μm
30 Ω	1.11mm	2.21mm	4.43mm
50 Ω	0.52mm	1.04mm	2.09mm
70 Ω	0.28mm	0.57mm	1.14mm

Table 2.1: Required trace widths for common impedance values. Top row is substrate (SiO_2) thickness. Left column is impedance value.

Lastly, trace width for the sensor has to be selected. For a SiO_2 substrate, the width of the trace is swept, and characteristic impedance results are calculated using an EM simulator, which can be found (Table 2.1). The top row shows substrate thickness, and the left column shows the impedance. Values inside the table are the required trace widths.

We had access to all the thickness values for the SiO_2 wafer. As an initial decision, 500 μm SiO_2 wafer is selected. For the trace width, 1mm is chosen to ensure impedance match. The coupling gaps are selected as 0.25mm for an initial guess. This is a common value used in many sources for coupling to microwave resonators. With all parameters decided upon, the resonator is illustrated in

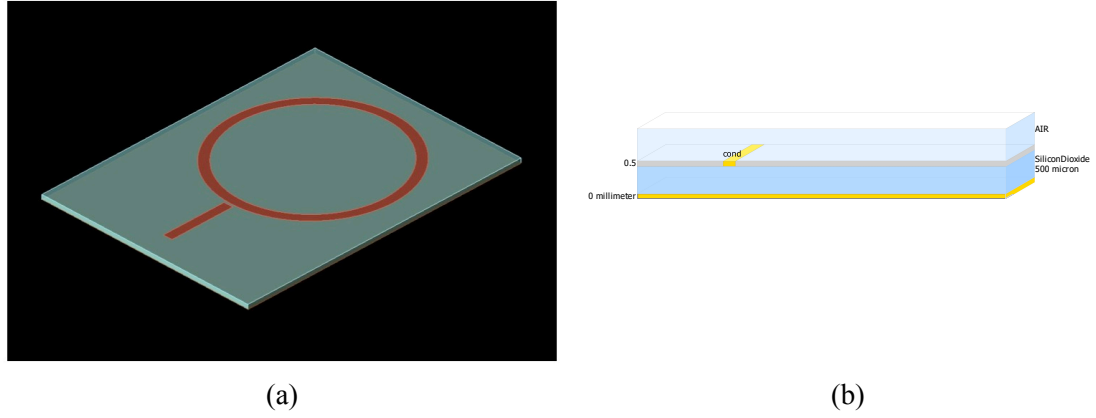


Figure 2.3: (a) Simple ring resonator with single feed line for S11 measurements and (b) substrate illustration with designed parameters.

Figure 2.3.a.

The s-parameter plot, calculated through EM simulator ADS Keysight, shows a solid agreement with our theoretical calculations. The resonance frequency of the first mode is at 2940 MHz, as opposed to our guess of 2700 MHz. Better estimates could have been done using more complicated design equations. On the other hand, it is seen in Figure 2.4 that the resonances are very narrow and weak. The resonance point is only -0.3dB deep for the first mode and similar for the second mode. These are the symptoms of under coupling.

The resonator is enclosed inside another ring to decrease loss and increase coupling, forming the Concentric Ring Resonator topology. Two feedlines are also added to the design for soldering SMA connectors. The resonant frequencies are shifted due to the different coupling method used. New resonance frequencies belonging to the outer ring are also introduced. On the other hand, the strength of resonant dips has increased drastically. S11 parameter is -3.92 dB at the resonance frequency, which is explained in detail in the next section.

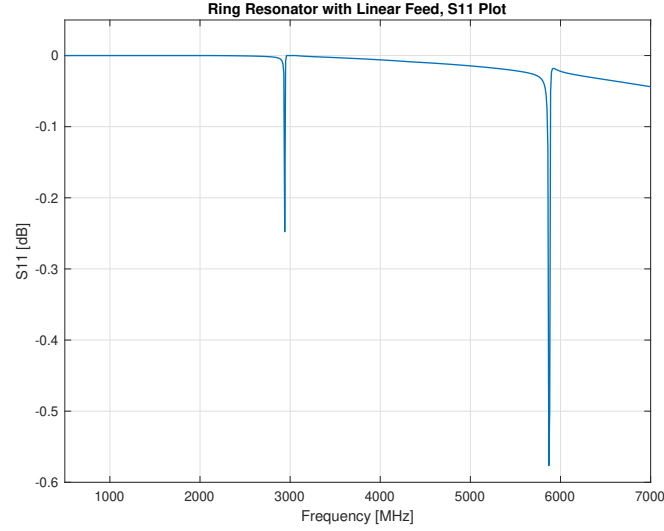


Figure 2.4: S11 plot for the ring resonator with linear feed. Notice the weak resonance dips, pointing to undercoupling.

2.1.4 Guided Design Process: Introducing the Split

A split has to be introduced onto the concentric ring resonator to form a sensing region with a cell-sized discontinuity. Electric fields are focused at the location of this split, creating a sensor structure.

Width of the split is chosen as $20\mu\text{m}$, matching the size of cells (MDA-MB-231) used during the experiment. Length of the split is selected as $80\mu\text{m}$, creating a suitable aspect ratio. Longer lengths can decrease the electric field focusing. Shorter lengths will shorten the event time, which will be explained in the experimentation section.

Trace width for the resonator was designed as 1mm. Width is lowered down to the split size through tapering to ensure a smooth transition of the EM waves. Most waves would have been reflected due to the sharp transition if the split was suddenly introduced without tapering. The split and tapering geometry is illustrated in Figure 2.6.

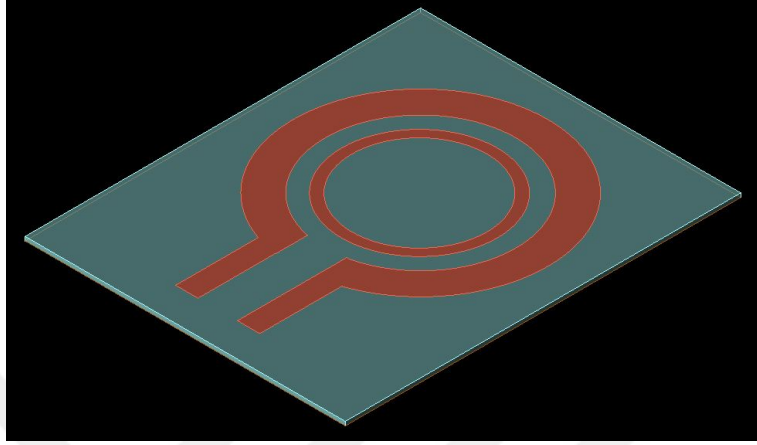


Figure 2.5: Concentric Ring Resonator geometry is formed. Inner ring is the resonator.

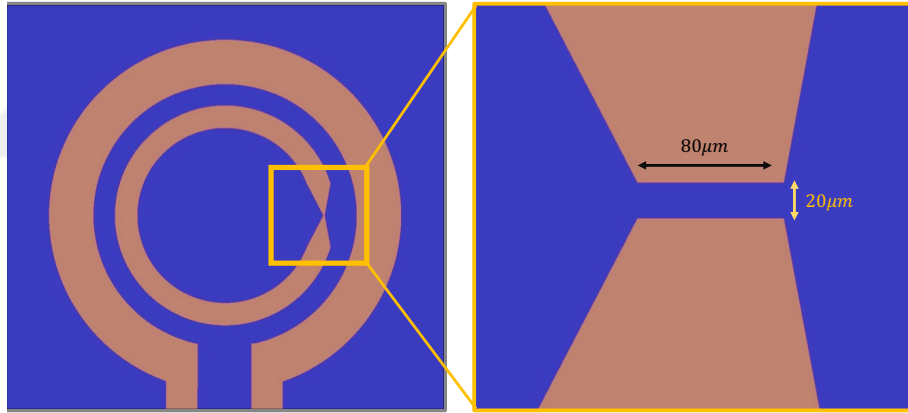


Figure 2.6: Split is introduced. Width of the resonator is brought down by tapering.

Before manufacturing, the design is simulated in COMSOL software to evaluate its performance. The gold layer, used for both signal and ground planes, is around 150nm thick. The thickness is bounded by UNAM cleanroom manufacturing capabilities. There is a three orders of magnitude difference between the thickness of the wafer and conductor layers. Therefore, the conductor layers are modeled with Perfect Electric Conductor (PEC) boundary condition, meaning the gold layer is infinitely thin and infinitely conductive. However, it does not mean that the impedance is zero for the layer. EM waves on the conductor are still subject to a finite impedance due to traveling in the substrate and air mediums. More information on this can be found from official COMSOL tutorial

posts [23] or texts on finite element method, with an emphasis on electromagnetism [24]. Simulation results for electric field intensity can be seen in Figure 2.7.a.

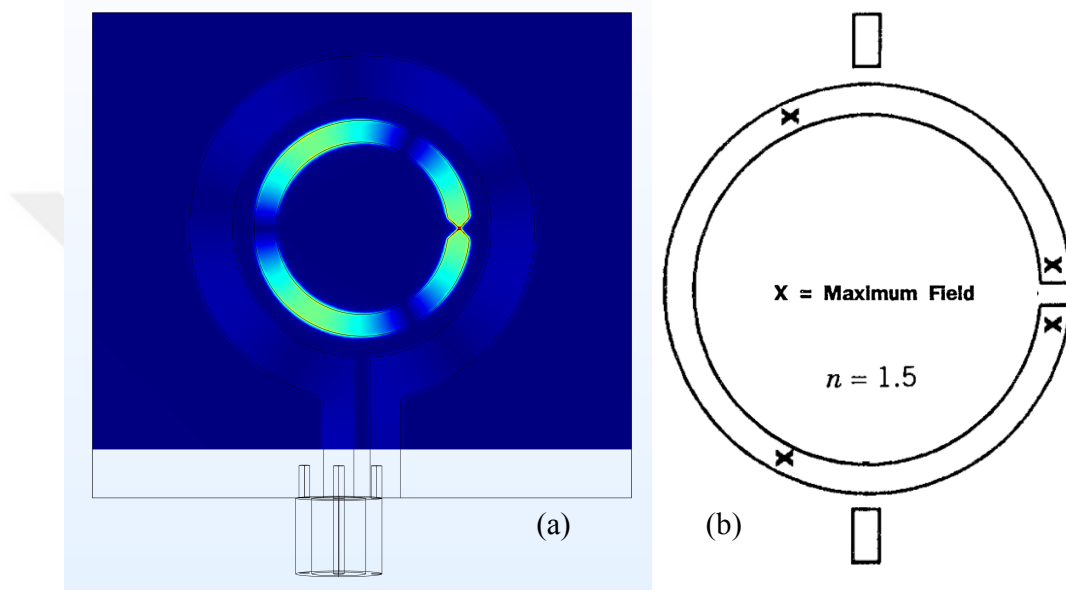


Figure 2.7: Comparison of (a) electric-field intensity figure from simulation and (b) theoretical electrical field taken from [2] (rotated 90 degrees for split locations to match).

The port is modeled as a lumped port with a wave excitation amplitude of 1V. The simulation is done from 1000 MHz to 7000 MHz, with 10 MHz steps. The maximum value of electric field intensity is searched through every frequency in this range, at every point on the geometry using COMSOL’s built-in function “Volume Maximum.” The maximum electric field intensity is at the sensing region, with an actual value of 70[kV/m], at 4390 MHz. One can notice that the locations of resonance points have also changed. Starting with our initial value of fundamental resonance at 2940 MHz, we expect to see the second mode ($n = 2$) at 5880 MHz. This new mode at 4390 MHz is called a “Split Resonant Mode,” discovered first by Wolff [25]. By calculation, it is seen that new modes emerge at half mode steps as seen below:

$$2940 \times 1.5 = 4410 \text{ MHz} \approx 4390 \text{ MHz}. \quad (2.5)$$

In addition, simulation results achieve a perfect agreement with the theoretical models shown in [2], as seen in Figure 2.7. Both from the (a) colored 2D graph and (b) drawing of the theoretical calculations, the electric field intensity is maximum at the split location. Using the split as the sensing region ensures the highest sensitivity for our sensor.

2.1.5 Guided Design Process: Improving the Design

Further improvements of the sensor are possible through an innovative modeling approach. Remember the introductory information provided in Section 2.1.1 about the series gap discontinuity. Capacitors C1 represented the fields fringing down to the ground layer. Our hypothesis is: by eliminating these fringing fields, the intensity of the electric field at the gap can be increased. This way, the sensitivity of the sensor can be enhanced.

Capacitors C1 can be eliminated by subtracting the ground plane from the design. Typically, this would result in an unusable geometry, as EM waves would not be able to propagate. However, the concentric split-ring resonator is unique because the ground plane is not essential for wave propagation.

In order to validate these claims, we rely on simulations. Both geometries, CSRR with and without a ground plane, are simulated using COMSOL 3D EM simulator. The ring topology (upper signal layer) is identical for both cases. The frequency range and the mesh settings are also identical. The only difference is that the first geometry has a ground plane, modeled through the usage of PEC boundary condition, whereas for the second geometry, the ground plane is omitted. For both simulations, boundaries of the enclosing air volume are modeled as perfectly matched layers. After simulations are done, the maximum value of electric field intensity is searched for all frequencies, through all points in the volume. This search is done by COMSOL's "Volume Maximum" built-in function. The resulting electric field is shown in Figure 2.8. The electric field intensity is focused at (and only at) the sensing gap.

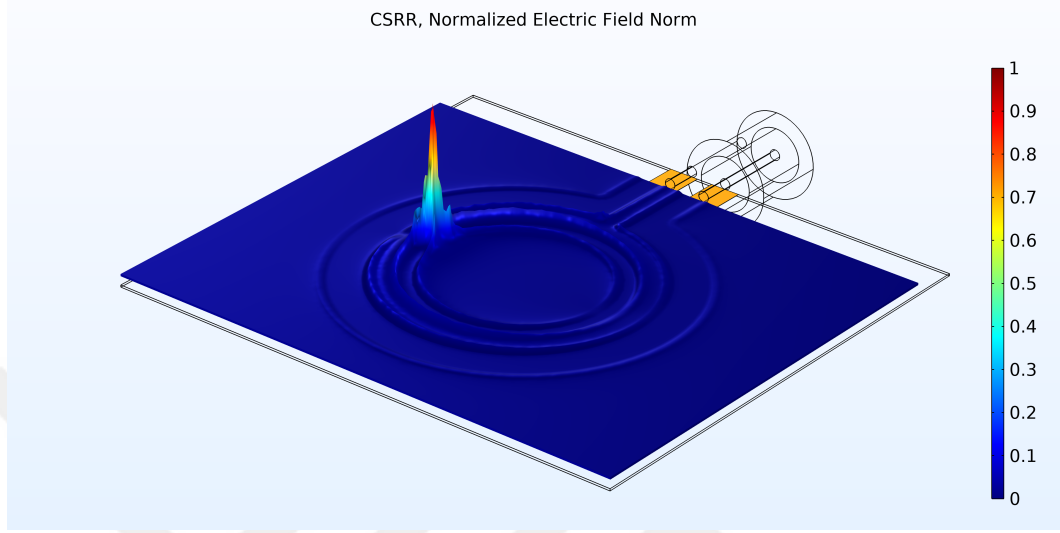


Figure 2.8: Illustration of the electric field intensity. The norm of intensity is mapped to z-axis to be more descriptive.

Even though the effects of focusing are clear from the figure, a quantitative comparison has to be made between two geometries. For the CSRR with ground plane, the maximum electric field intensity is $70[\text{kV/m}]$. On the other hand, the maximum electric field intensity for the CSRR without ground plane is $163[\text{kV/m}]$. This is more than twice stronger than the ground plane topology. An illustrative comparison can be found in Figure 2.9.

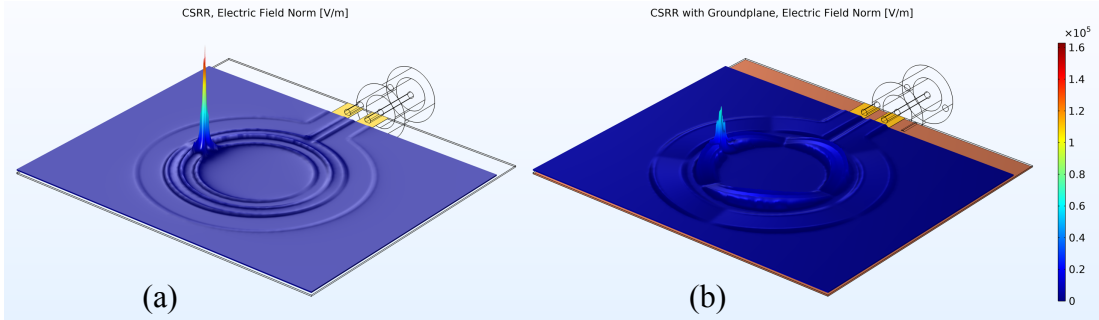


Figure 2.9: Comparison of the electric field intensity for (a) CSRR without ground plane and (b) CSRR with ground plane. The same color scale is used for both. CSRR without ground plane has a stronger and more focused electric field intensity distribution.

With these observations, the sensor's framework is established. The next chapter explains the fabrication process for the CSRR without ground, as it will be

our sensor of choice for the following chapters.

2.2 Device Fabrication

The wafer used is Q100020ED-O from Quartz Unlimited company. It is a dual side polished, low-loss wafer with the specifications:

- Relative Dielectric Constant: 3.75
- Loss Tangent: 0.0012
- Electrical Resistivity $> 10^{18} \Omega.m$

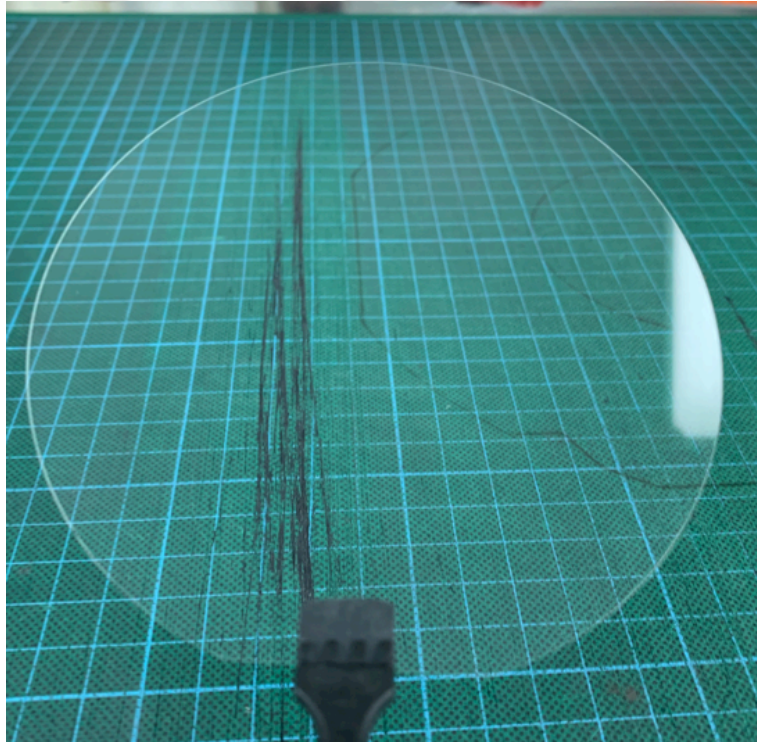


Figure 2.10: Wafer Q100020ED-O from Quartz Unlimited company.

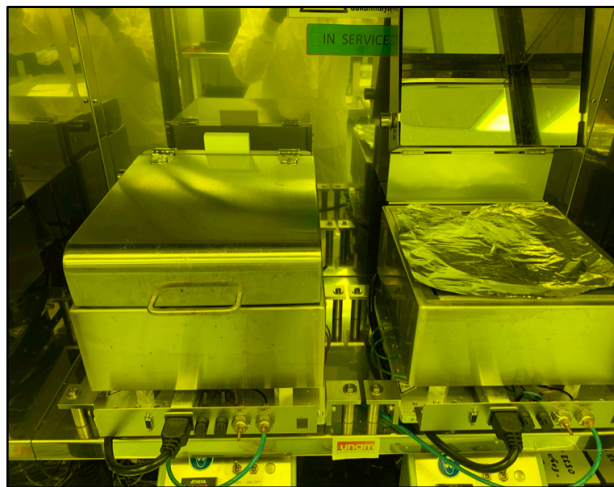
Fabrication is done in the UNAM cleanroom. The process begins with the dicing saw machinery. It is used to cut the wafer into two equal parts. Parameters for dicing saw are as follows:

- Work Shape and Size: Round 120mm
- Work Thickness: 0.500 mm
- Tape Thickness: 0.090 mm
- Feed Speed Level: 3
- Spindle Rev.: 30000 rpm

The next step is photoresist covering. Before the process, the surface of the wafer is cleaned using acetone followed by isopropyl alcohol. All is dried with pressurized nitrogen. The surface of the wafer is covered with photoresist, using a plastic pipette. Spin coater is used for this process. The coated photoresist is AZ5214, which is a widely used positive photoresist.



(a)



(b)

Figure 2.11: Diced wafer is covered with photoresist. Coating is done in (a) spin coater. Afterwards, the wafer is (b) baked in the ovens.

The parameters for spin coater dictate the end result of the resist thickness.

- Spin Duration: 30 seconds
- Spin Velocity: 4000 rpm

- Acceleration: 1000 rpm/s

With the parameters above, the resist thickness comes out as $1.5\mu\text{m}$. After the coating process, the photoresist is baked at 110°C for 60 seconds, using the ovens seen in Figure 2.11.b.

The photolithography (exposure) step comes up next. Lithography mask, as seen in Figure 2.12.b, is inserted first. After inserting the wafer, the lithography pattern is aligned to be centered on the diced wafer.

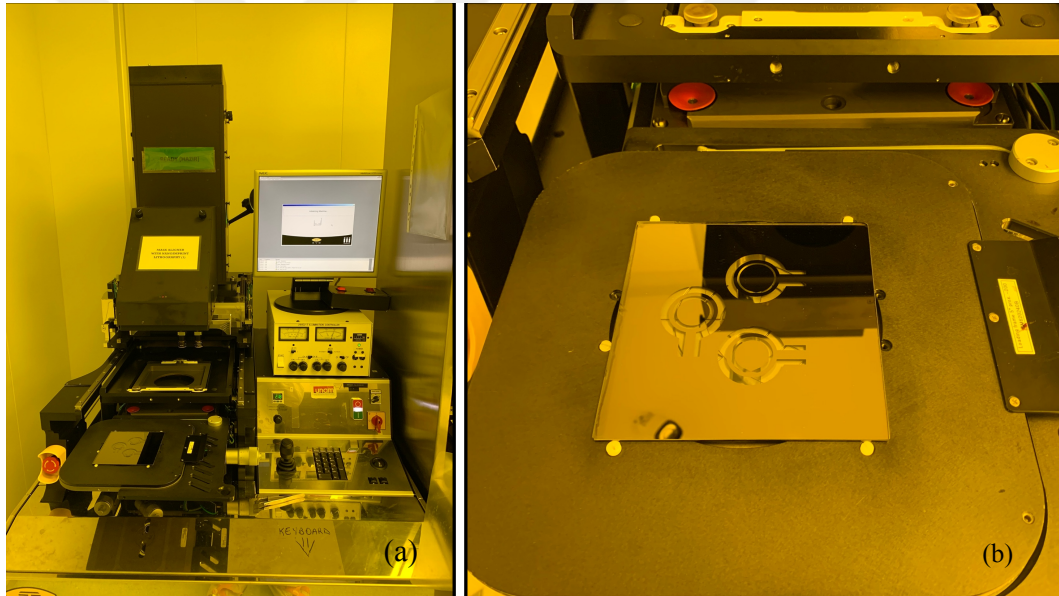


Figure 2.12: (a) Photolithography machine and (b) mask for controlling exposure geometry.

Parameters for exposure are listed as:

- Exposure Mode: Constant Dose
- Contact Mode: Soft Contact
- Separation: 50 μm
- Mask Thickness: 2.4 mm
- Substrate Thickness: 0.25 mm

- Resist Thickness: $1.5\mu\text{m}$

After the exposure process, a developer solution is prepared. For the developer solution, one part of AZ400K is mixed with four parts (volumetric) of water. The wafer is dipped into the developer solution for 60 seconds. The developing process is terminated by dipping the wafer into deionized water for 30 seconds.

Finally, thermal evaporator is used to coat the conductor layer of the CSRR. After inserting the wafer in place, the coating process begins. First, a thin layer (7 nm) of chromium is deposited onto the wafer. 200nm of gold is deposited afterwards, and the process is completed. The wafer is put into a pool of acetone for overnight lift-off.



Figure 2.13: Photo of the final CSRR device on diced quartz wafer.

The device after the microfluidic channel integration and SMA connection is

illustrated in Figure 2.14. The procedure for microfluidic channel integration is covered in Chapter 3.4 because it is identical for both devices.

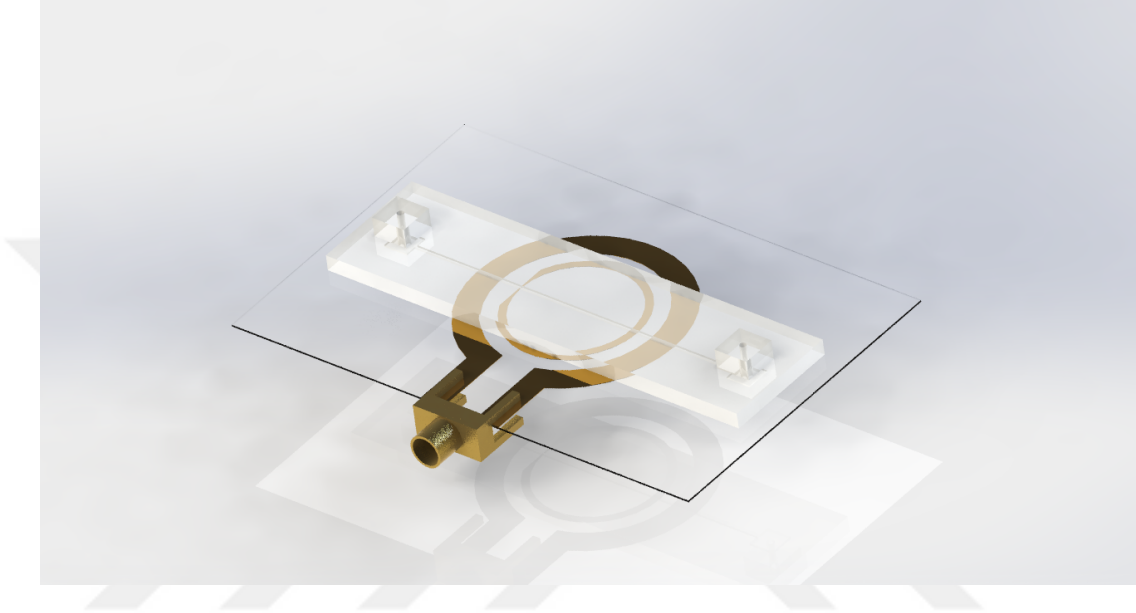


Figure 2.14: Illustration of the final CSRR device on diced quartz wafer, with integrated microfluidic channel and soldered SMA port.

2.3 Experimental Procedure

The experiment is based on passing single cells (or microparticles) through the integrated microfluidic channel and recording the change in resonance frequency. Therefore, the setup has two major parts: electronic measurement circuitry and microfluidics setup.

The microfluidic setup consists of a pressure controller, Fluigent MFCS, used to pressurize liquid into the microfluidic PDMS channel. Inside this liquid are the cells (MDA-MB-231) or microparticles (Sigma-Aldrich $20\mu\text{m}$ Polystyrene). Fluigent MFCS is specially selected for our experiments, considering it does not include any moving parts. Syringe or piston pumps are unsuitable for usage in the micro-liter range because they introduce pulses during flow. These flow pulses, through the sensor, can translate into frequency pulses in our electronic

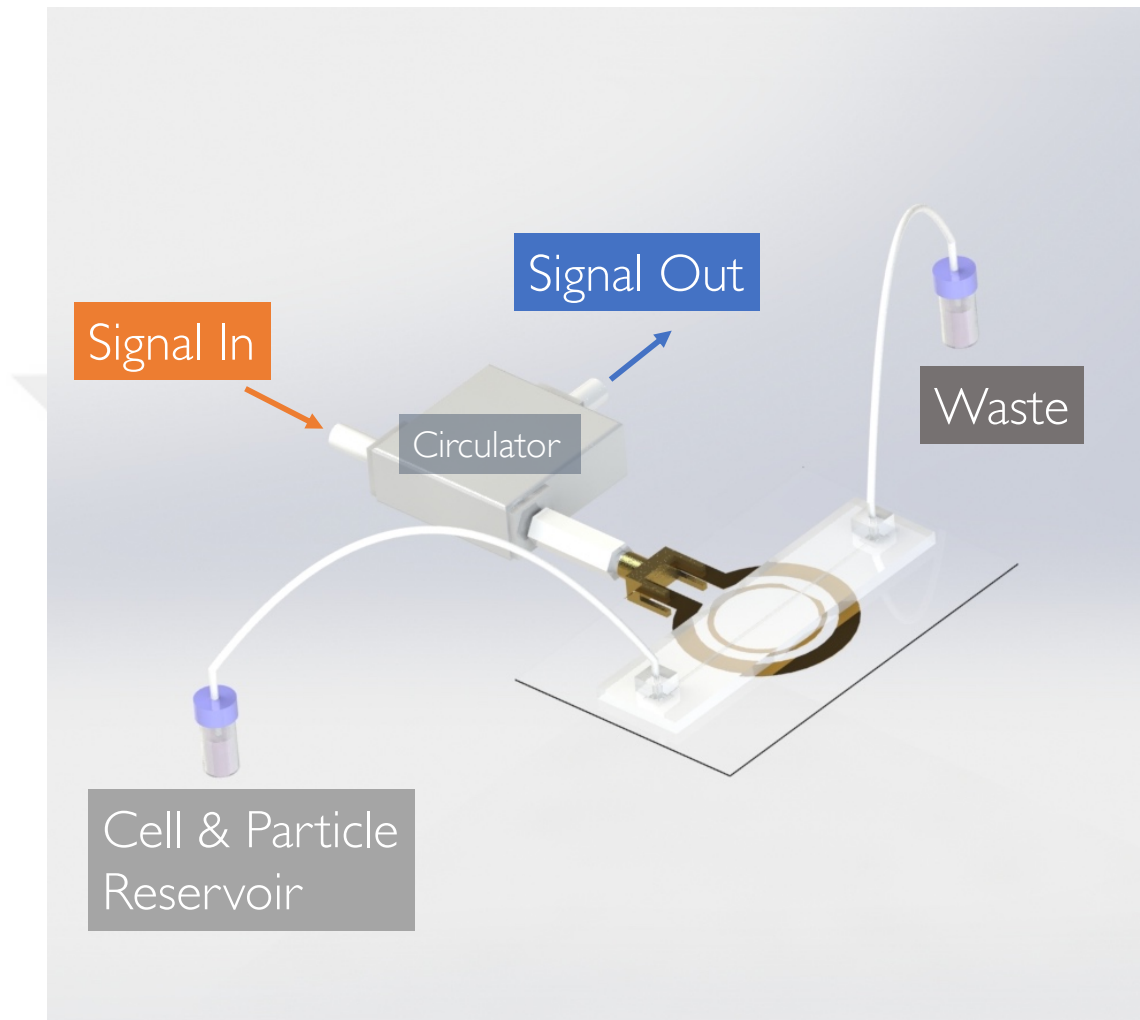


Figure 2.15: Experimental setup illustration.

measurements. Fluigent MFCS prevents this using a different pump technology. It can ensure steady fluid flow by driving the flow at constant pressure.

Electronic measurement setup, responsible for driving our circuit at the resonance frequency, consists of three major components:

- Signal Generator
- PI Controller
- Phase Detector (blue circle)

These components can be seen in Figure 2.16. The figure is deeply simplified for clarity. The signal generator is SMB100A with a frequency range of 100kHz to 20 GHz. It is governed by the PI controller, implemented digitally in PC LabView software. The phase detector is a lock-in amplifier from Zurich Instruments, MFLI series. The operation frequency of the MFLI device is limited to 5 MHz. For microwave domain operation, upconverter and downconverter circuits are used. Details of the measurement setup are covered in [19].

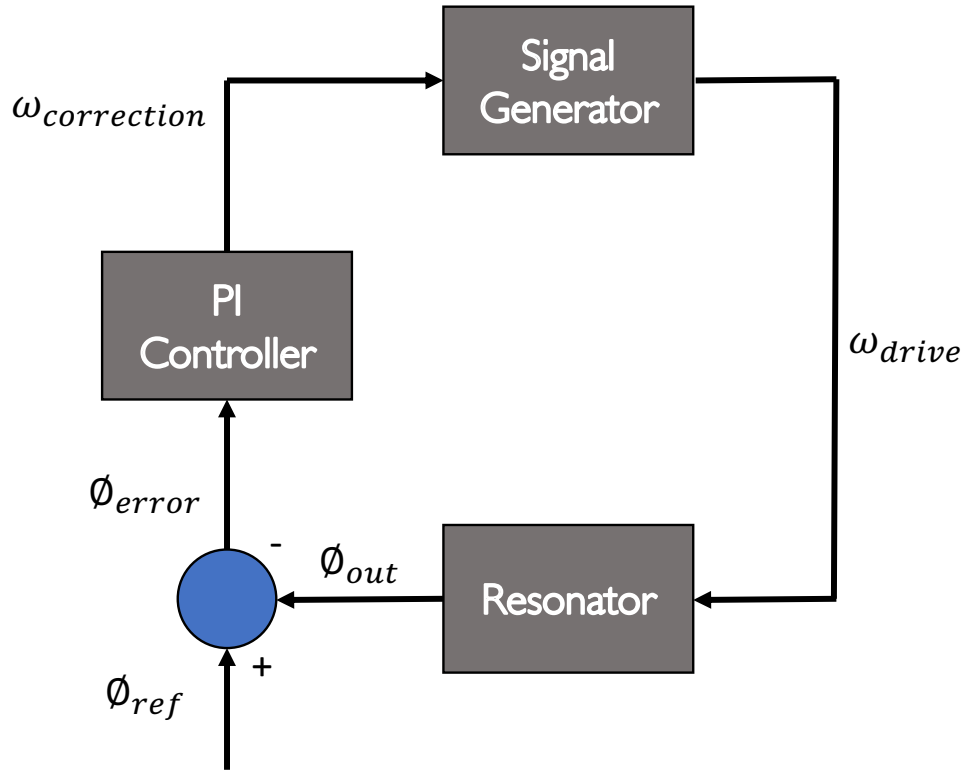


Figure 2.16: Electronic measurement setup schematic.

CSRR sensor is a single-port device. For experimentation, the sensor needs to be driven, and information should be read from it simultaneously. Therefore, a circulator is used. After using the circulator, the S21 parameter is measured using a vector network analyser. Typical S11 plot for CSRR can be seen in Figure 2.17. Only a single one of these resonance dips survive after connecting the circulator.

The reason is that the circulators have a narrow bandwidth. For example, PE8401 is the circulator used during the experiments. It has an operating bandwidth of 2-4 GHz. Therefore, only the second dip survives after connecting the circulator.

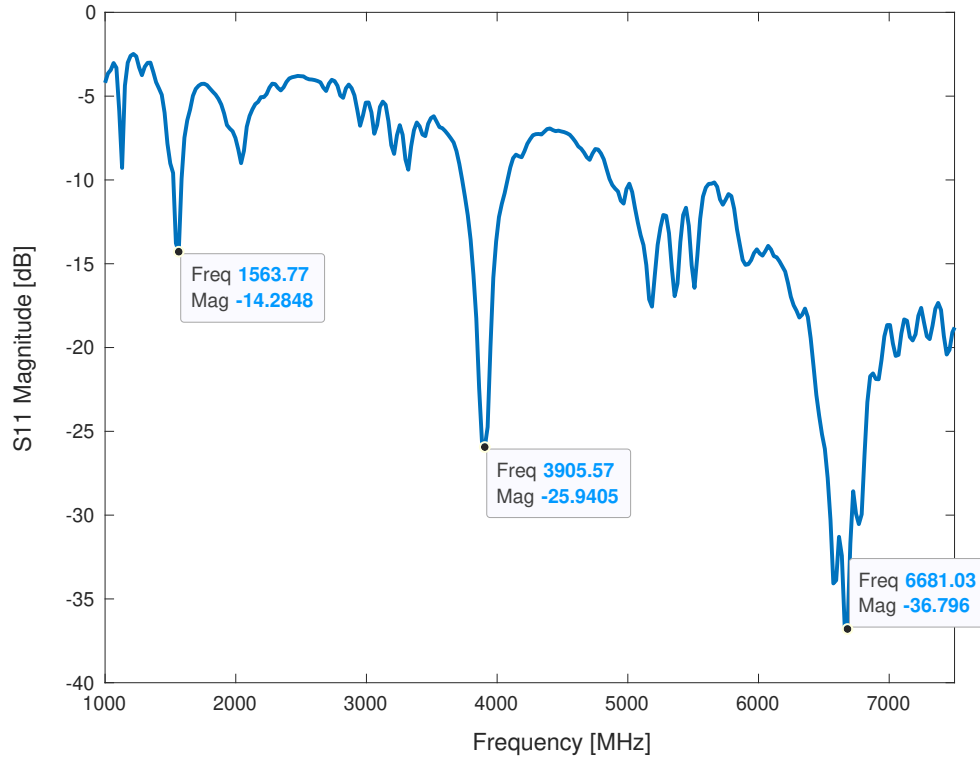


Figure 2.17: Typical S11 response plot for CSRR sensors.

After the resonance location is identified, the driving circuit is set to pursue this frequency. The circuit also requires the user to set PI controller parameters. As cells pass, the resonance point changes. Shifts in resonant frequency are recorded. All of the events are visually validated by a precision microscope.

2.4 Experimental Results

Two experiments were performed: polystyrene and single-cell experiments. The polystyrene experiments are valuable as calibration experiments because particle size, shape, and material are closely controlled. Single-cell experiments are an

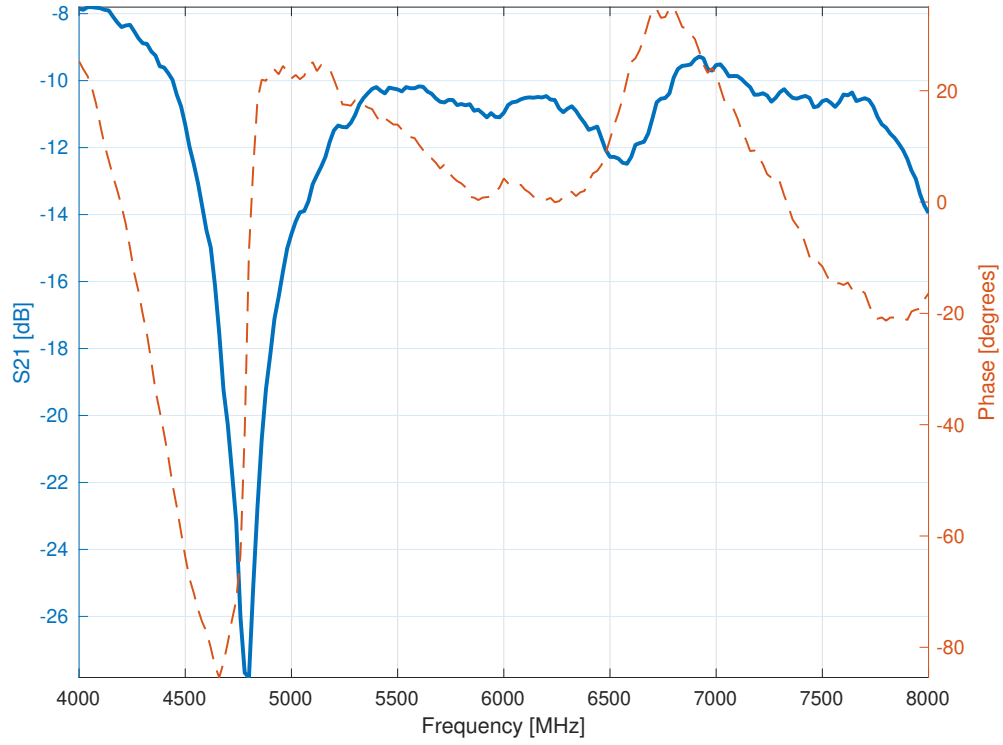


Figure 2.18: Typical S21 response plot for CSRR sensors, after circulator integration.

opportunity to evaluate real-life performance of our sensor, measuring its performance with actual cells. Unlike polystyrene particles, cells have no definitive shape. The variance of their size is large compared to the variance between polystyrene particles. Also, the material inside them can differ from cell to cell.

Events are in the form of peaks. To better explain the process, the illustration in Figure 2.19 can be helpful. As cells pass, the peak is formed on the resonator frequency vs. time graph. The quality of the peak is dependent on PI controller parameters.

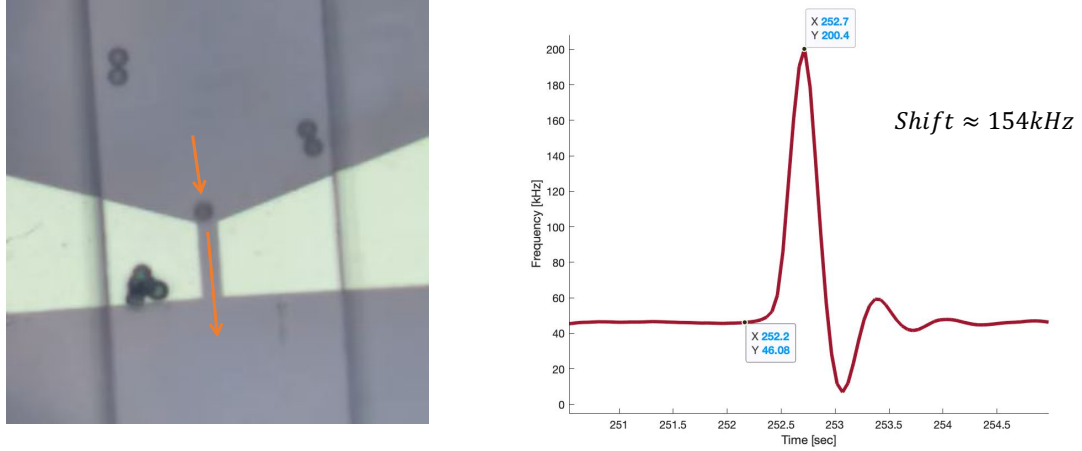


Figure 2.19: Illustration of a typical event, both in visual and electrical measurement domains. Particles are $20\mu\text{m}$ polystyrene microparticles.

2.4.1 Polystyrene Experiments

Particles have a mean diameter of $20.0\mu\text{m}$. Sigma-Aldrich provides a Certificate of Analysis for these microparticles, specifically issued for our batch. According to this document, the polystyrene particles have the following specifications:

- Particle Composition: Polystyrene
- Nominal Particle Diameter: $20.0\mu\text{m}$
- Standard Deviation: $0.30\mu\text{m}$
- Measuring Technique: Coulter Multisizer III

For the experiment, over 40 events are collected. Histogram of the frequency shifts induced by $20.0\mu\text{m}$ polystyrene particles can be seen in Figure 2.20.

The deviation of the shifts is caused by positional dependency. Not all particles pass precisely over the sensing gap. There are slight variations. These variations cause differences in the induced frequency shifts. The problem is discussed, and solutions are proposed in further sections.

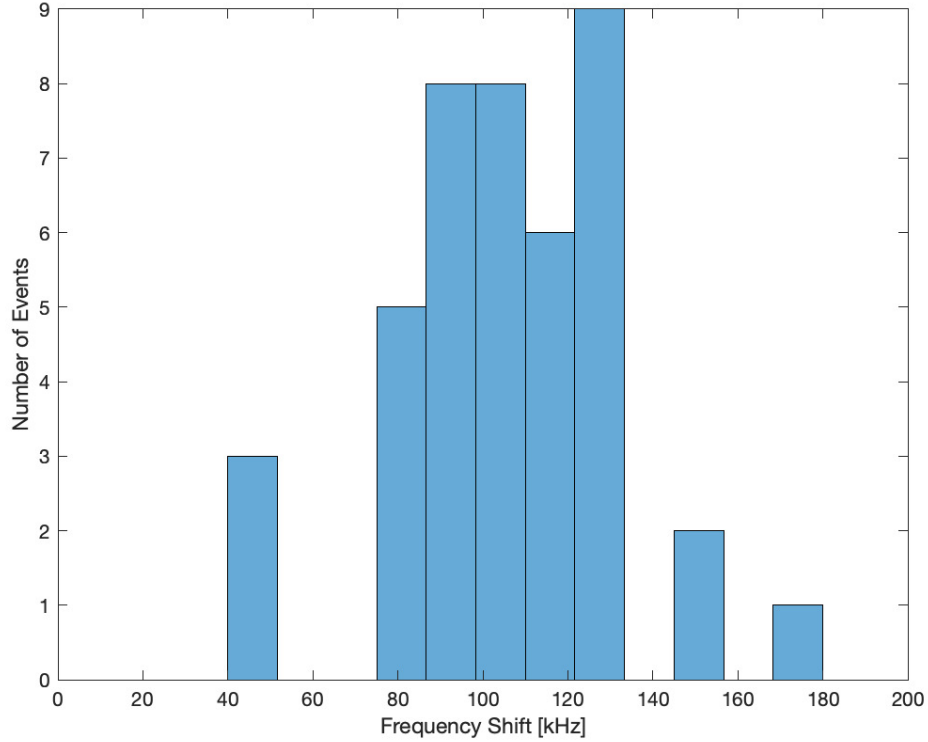


Figure 2.20: Histogram of the frequency shifts for 20.0 μm polystyrene experiment.

2.4.2 Single Cell Experiments

The cell line used is MDA-MB-231, purchased from ATCC (Manassas, Virginia). In order to make validation simpler and more accurate, cells are labeled with a green fluorescent protein. This is used to distinguish cells optically.

CSRR sensors used for single-cell experiments are identical to the ones used for polystyrene experiments. There are various experiments done with MDA-MB-231 cells. Even though the cell line is identical, all cells are at different life stages, with different amounts of biological material. Therefore, their sizes vary throughout the experiment for different cells. Due to this, the response we get in the signal domain also changes with time.

In order to eliminate this effect and demonstrate the sensor's true capabilities,

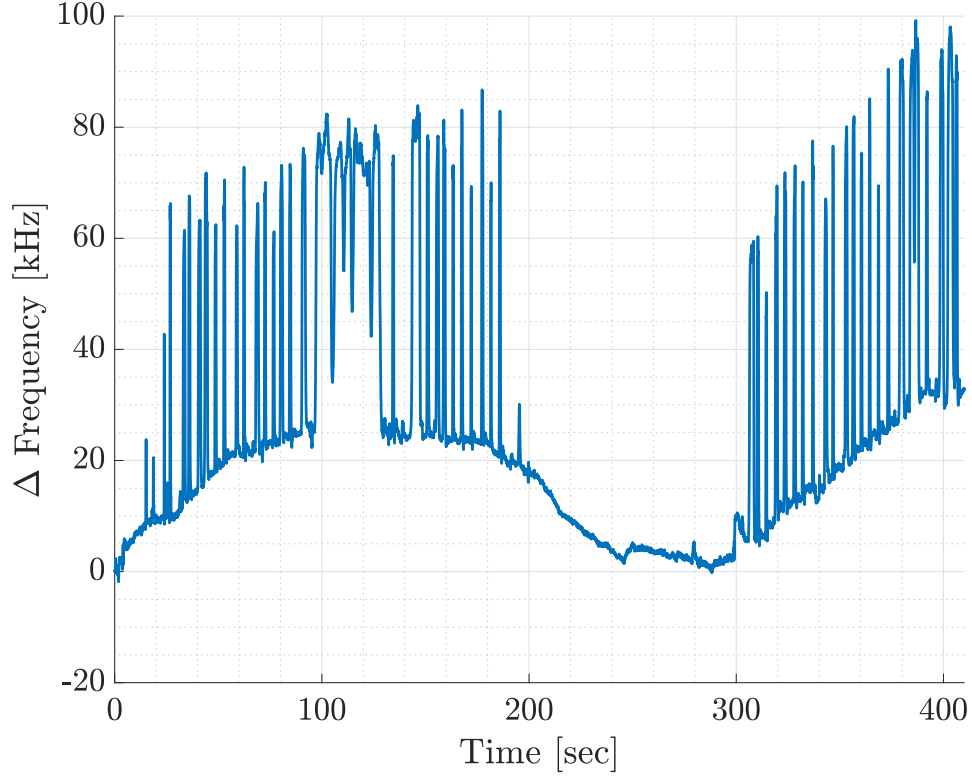


Figure 2.21: Time domain data for the MDA-MB-231 experiment.

it is believed that the data for a single identical cell would be more informative for the reader. For the experiment, a single cell is targeted and trapped in the channel. Then it is shifted back and forth many (more than 50) times. The frequency shifts in the time domain can be observed in Figure 2.21.

Histogram of the results is shown in Figure 2.22. All of the different frequency shifts are consecutive and belong to a single cell. Normally, it would be expected for the sensor to give identical responses to identical cell passage events. The variance in the results is due to positional sensitivity. Even though the same cell is responsible for all the events, the position changes slightly with different events. The channel is $40.0\mu\text{m}$ wide, whereas the cells are approximately $20.0\mu\text{m}$ in diameter, on average.

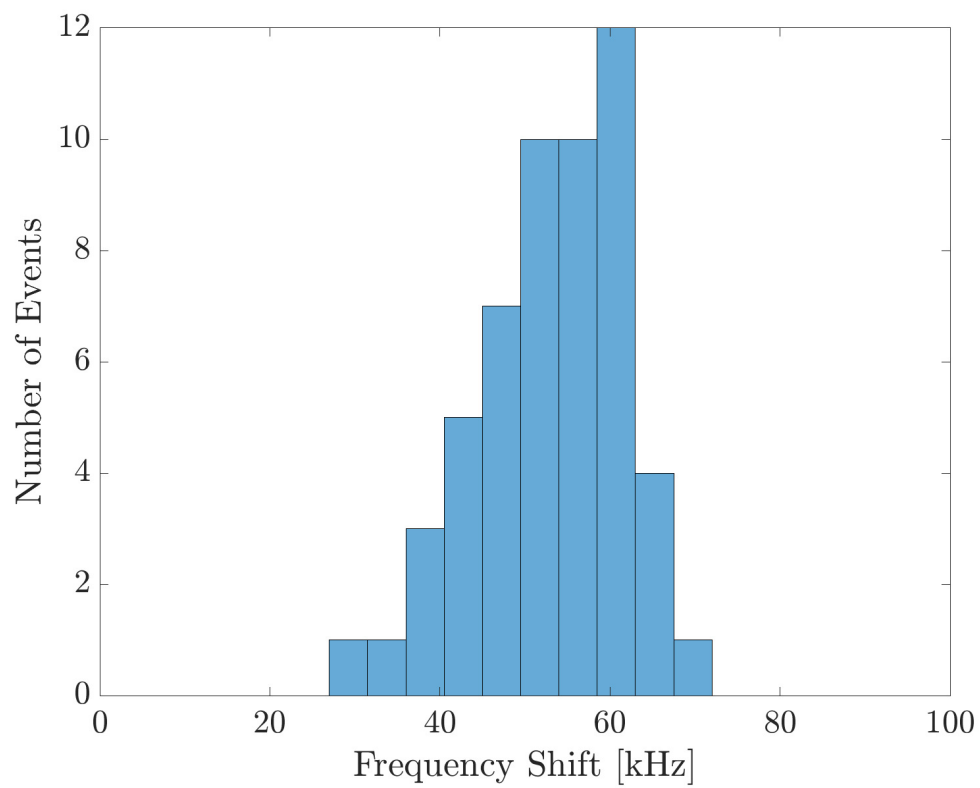


Figure 2.22: Histogram of the frequency shifts for the MDA-MB-231 experiment.

Chapter 3

Micro-particle Detection with Standalone SRR Integrated Oscillators

After the achievements made with the concentric split-ring resonators, there was a drive to improve the method further. The main problems with our experimental setup were:

- It was slow compared to usual electronic measurement rates, limited by 20 samples/s due to communication speed between the devices.
- Success of the results was dependent on PI controller parameters. Each experiment required setting of parameters by the experimenter.
- System required highly sophisticated instruments to operate. It was not cost-effective.

In addition to these quantitative points, there is also a qualitative improvement. With the PLL system, the resonant frequency is “tracked” continuously. It would be more desirable if no tracking were necessary, and the resonator’s resonant frequency was handed us without any additional effort.

With these ideas in mind, the standalone oscillator idea is considered. The new oscillator design will require no sophisticated instruments to track its resonant frequency. Therefore, the delays due to communication overhead between these instruments will be eliminated. Lastly, the success of the results will not depend on the correct setting of some parameters.

3.1 Theory of Oscillators

In most basic terms, the act of designing an oscillator means composing a circuit that satisfies Barkhausen's Criteria for oscillation. There are various concerns regarding this criteria, majorly focused on the idea that it can yield nonoptimal designs. While some of these concerns might be true, the method is still prevalent among oscillator designers. Barkhausen's Criteria for oscillation, determined from the open-loop frequency response, are as follows [26]:

- (i) Frequency of the phase zero crossing is the oscillation frequency.
- (ii) Initial gain at oscillation frequency must be greater than 0 dB.

When these necessary conditions are satisfied and the loop is closed, a signal begins building up. Amplitude increases to a level where nonlinearity effects begin to emerge and limit the amplitude. At steady-state operation, the gain becomes unity, and oscillations are sustained.

There are various sources [27, 28, 29, 30] that establish the theoretical framework for oscillators through rigorous mathematics. On the other hand, almost all of these sources at least have a chapter where they introduce (and suggest) using computer-aided design software, especially for designing at higher frequencies. Components, both passive or active, behave differently at higher frequencies than in lower frequencies. Caring for all of these new effects during the design process can become overwhelming for the designer. Therefore, computer-aided design is used for the design of oscillators.

3.2 Design of SRR Oscillator

In Chapter 2, concentric split-ring resonator was proven as a promising sensor structure. However, the same resonator geometry could not be used for oscillator setup because the resonance consisted of dips as shown in Figure 2.17. For oscillators, the resonance has to be in the form of a peak. Therefore, a resonator with a peak resonant behavior is designed.

For satisfying Barkhausen's Criteria, an amplifier section is required. Using off-the-shelf amplifier components was not optimal due to associated high noise figures. Therefore, a single-transistor amplifier section is designed from scratch with much better characteristics in a narrow bandwidth.

Lastly, resonator and amplifier sections are connected together. The “integration” had to be made with care because there were no additional components for satisfying the phase shift criterion. Trace lengths and coupling strengths are used for satisfying this criterion without external components. With this, the design is finalized.

In the previous chapter, it was shown that it is possible to design a resonator without a ground plane. However, this time every component is designed from scratch, and that is already challenging. To eliminate the ground plane from the resonator and integrating it to the rest of the circuit was unnecessarily complicated as the first step of an experimental project. Therefore in this chapter, the bulk of the oscillator geometries use a conventional microstrip split-ring resonator. Experiments with resonators omitting ground planes are also performed but require improvements.

3.2.1 Designing the Split-Ring Resonator

Details regarding the design process of a resonator are shown in Chapter 2.1.3. In order to prevent the text from becoming repetitive, most of the details are

omitted in this section.

We will begin with a rough figure of the diameter of the split-ring resonator. This time it is desired for the oscillator to operate at 2 GHz. The reason for this is to keep dielectric losses low. Unlike the CSRR sensor, the oscillator project has cost-effectiveness as one of its requirements for success. Therefore the substrate material used is classic FR4, the most popular substrate for PCB production. According to the manufacturer (JLCPCB Company), it has the following specifications:

- Material: FR4 (Shengyi S1141)
- Relative Dielectric Permittivity: 4.5
- Loss Tangent: 0.0200

Within these parameters, loss tangent requires most of the attention. It is more than 16 times larger compared to SiO_2 substrate. The losses increase even more as frequencies get higher. Therefore, the operation frequency is kept as low as possible. In addition to this, designing a circuit with $50\ \Omega$ characteristic impedance on an FR4 substrate requires a large trace width. Table 3.1 shows a set of results from the ADS Keysight EM simulator, computed for common characteristic impedance values for different FR4 thicknesses.

FR4 Thickness	Impedance	Trace Width
1.00 mm	30 Ω	4.11 mm
	50 Ω	1.90 mm
	85 Ω	0.64 mm
1.20 mm	30 Ω	4.95 mm
	50 Ω	2.29 mm
	85 Ω	0.78 mm
1.60 mm	30 Ω	6.61 mm
	50 Ω	3.10 mm
	85 Ω	1.05 mm

Table 3.1: Required trace widths for representative (low-mid-high) impedance values on FR4 board with various thicknesses.

A substrate thickness of 1mm is chosen because achieving low characteristic impedance is challenging with thicker substrates. Trace width of 1mm is chosen, yielding a characteristic impedance of approximately 70Ω . Using the built-in FR4 substrate model in ADS Keysight software, the required diameter for the SRR sensor is found as 26mm. The corresponding s-parameter plot can be found in Figure 3.1.

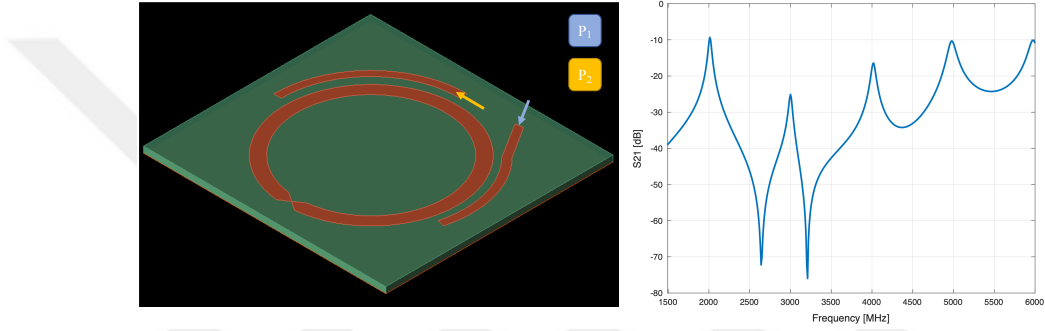


Figure 3.1: Designed Split-Ring Resonator for oscillator integration and corresponding s-parameter plot.

The resonant peak at 2 GHz is clear and significant. Quality factor is calculated by Equation 3.1. It yields a value of approximately 53, which can also be visually validated using Figure 3.2. Numerical calculation for the quality factor is as follows:

$$Q_{-3dB} = \frac{2.012 \text{ GHz}}{(2.030 - 1.992) \text{ GHz}} = 52.95. \quad (3.1)$$

Even though there are other peaks in the s-parameter plot, the oscillator will be designed to oscillate at the 2 GHz peak. Gain at higher-order modes will decay as the transistor does not have enough gain to supply oscillations at those frequencies.

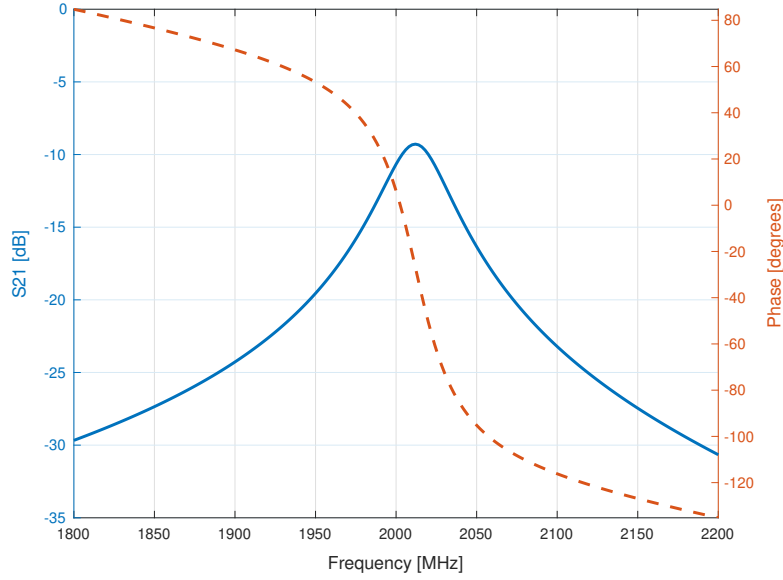


Figure 3.2: Zoomed version of the resonant peak located at 2GHz. Phase response is also illustrated.

3.2.2 The Design of the Amplifier Section

Before the design of the amplifier, experiments using an off-the-shelf amplifier are performed. However, these amplifiers are not suitable for oscillator operation because they are not designed for this purpose. As an example, one of the oscillators experimented with is the ZX60-43-S+ from Minicircuits company. It has a frequency range from 0.5 to 4000 MHz. On the other hand, the noise figure is 5.4dB typical, which is too much for our purposes.

Therefore, the design is carried out from scratch, starting by selecting a low-noise RF transistor. One of the best candidates is BFP740 from Infineon Technologies. This bipolar transistor is designed particularly for microwave frequency operation. In addition to this, Infineon Technologies company provides a complete EM model for each of its transistors, which becomes extremely useful during the design process.

Single transistor common emitter amplifier topology is chosen for the design process. The biasing is calculated using the design rules found in [31]. The values

of resistors are rounded to the nearest standard lumped component values. Three resistors are used in the design. Resistors R1 and R2 form a stiff voltage divider for biasing the transistor. Rc is the collector resistor for dictating collector current. Setup for the biasing simulation can be seen in Figure 3.3.

Biasing resistors could have been chosen with higher values. This would form a stiffer voltage divider. On the other hand, the equation for thermal noise is:

$$V_{Johnson}(rms) = \sqrt{4kTRB} \quad V(rms). \quad (3.2)$$

Johnson noise is proportional to the square root of resistor value [32]. Noise generated at the biasing circuit will get amplified and turned into a much more significant noise component at the output section. It will also translate into phase noise for the final oscillator. Therefore, resistor values are chosen as 1000Ω for R1 and 200Ω for R2. The collector resistor has a resistance of 50Ω . With these values, the base is biased at 820 mV, assuming a 5V supply voltage.

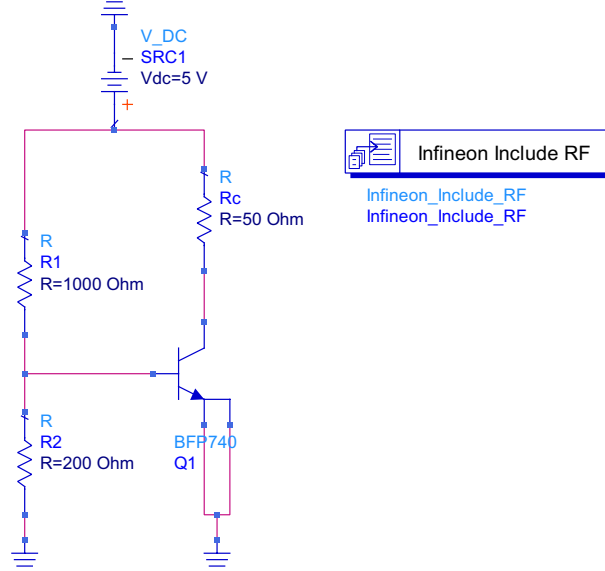


Figure 3.3: Bias scheme for BFP740 transistor.

The stability of the transistor amplifier is checked with K-Factor [33] test. It was found stable, but there was room for improvement. It is advantageous to have some safety factor for the stability during the typical design process of an amplifier. However, this improvement is not made for our amplifier. Additional

components may be required if the stability is checked beforehand. On the other hand, the oscillator design is not finalized yet. By the addition of the resonator section, the stability can be improved without any extra components. Therefore, only with these component values, the design for the amplifier stage is concluded.

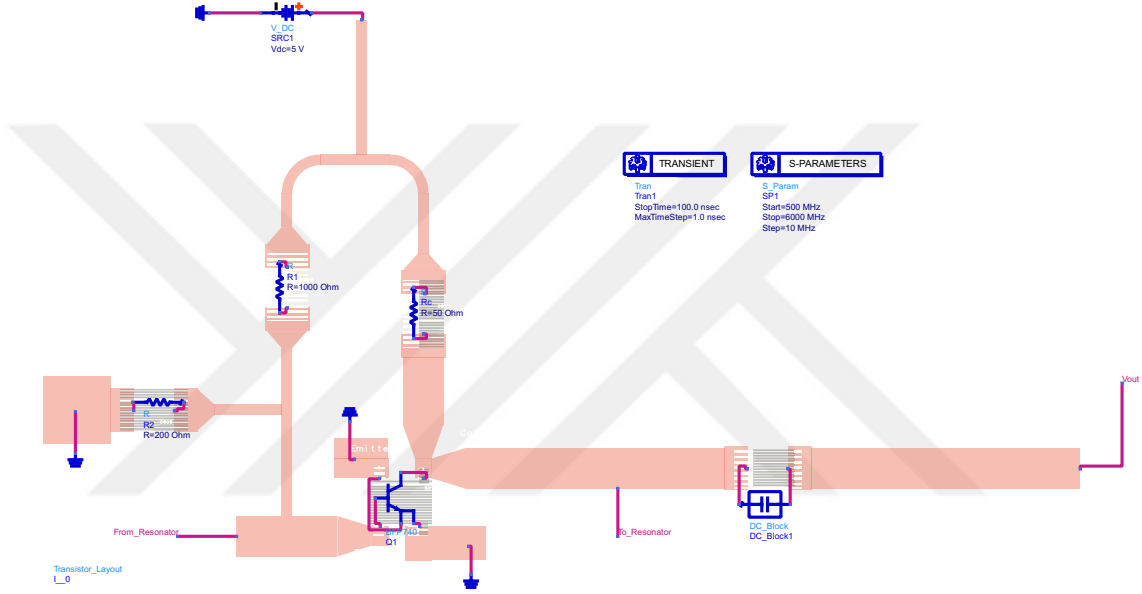


Figure 3.4: Layout for the amplifier with calculated component values.

Layout for the amplifier is illustrated in Figure 3.4. The signal from the collector port is extracted to be fed into the resonator section. This signal is denoted by the “ToResonator” label. Feedback from the resonator is fed into the base port. This signal is denoted by the “FromResonator” label. The resonator will be integrated with the amplifier using these ports.

3.2.3 Connecting the Subsections and Finalizing the Design

Amplifier and resonator sections are connected together with carefully designed transmission lines and coupling gaps. On the other hand, considering the complicated nature of designing an EM project at this scale, experimenting with various configurations is essential to converge to a successful design.

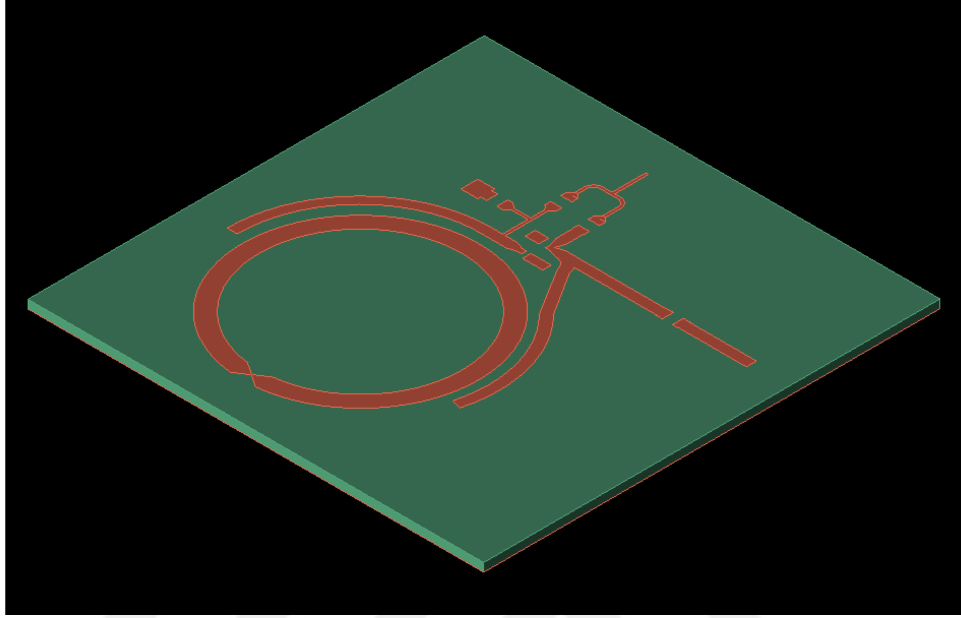


Figure 3.5: Complete layout for the oscillator.

Notice that in Figure 3.5 various trace widths are used throughout the design. For example, transmission lines for coupling to the resonator are much thicker than those connected to the voltage supply. A straightforward comparison can also be observed in Figure 3.4. At DC, the width of a transmission line has no effect on small-signal propagation. However, high-frequency waves such as the oscillations at 2GHz are incapable of propagating on these thin traces. Therefore the biasing circuit, which consists of thin transmission lines, does not allow high-frequency signal propagation. This way, the DC supply and the high-frequency oscillator sections are partially isolated from each other.

The design is finalized, and the response of the oscillator is measured at the output port for verification. From Figure 3.6, it is seen that the Barkhausen Criteria for oscillation are satisfied at approximately 2 GHz. The gain margin at the frequency of oscillation is 10dB.

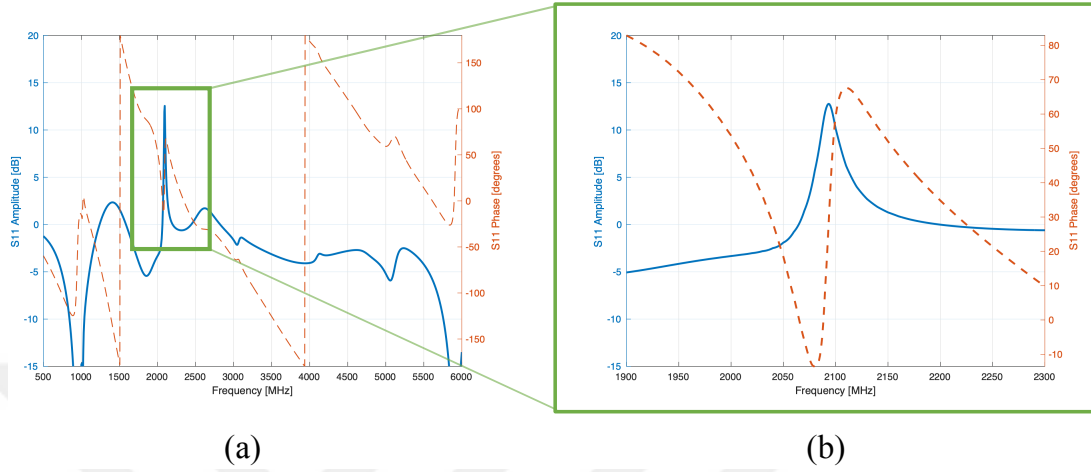


Figure 3.6: S-parameter plot for the oscillator, probed by the Vout port. At 2 GHz, phase curve has a zero crossing and the gain is adequate. Barkhausen Criteria is satisfied.

3.3 Manufacturing: Main Oscillator Boards (Version 4.1)

Designs are exported to Altium Designer software for Gerber File generation. Multiple via connections are added for low-impedance ground contact, especially at emitter traces. Solder mask is omitted from the design because its EM characteristics at microwave frequencies are not well documented. It is also known to be a lossy material at high frequencies.

Sensing gap width is also increased to $120\mu\text{m}$ due to manufacturing capability limitations provided by the manufacturer. The conductor surface finish is ENIG-RoHS to reduce surface roughness. The resulting product after soldered components can be seen in Figure 3.7.



Figure 3.7: (a) Oscillator layout designed in Altium Designer, and (b) manufactured design.

3.4 Microfluidic Channel Manufacturing and Integration

Polydimethylsiloxane (PDMS) material is used for microfluidic channel manufacturing. The process consists of 5 steps. In the first step, a silicon wafer is used to prepare the mold. On this wafer, the microfluidic channel pattern is created using SU-8 material. Steps are nearly identical to the photolithography stage found in Section 2.2, with spin coating, baking, exposure, and developing. Further details for this process can be found in [19].

After PDMS is cured and peeled from the mold, the next step is to bond it to the FR4 substrate. Commonly a SiO_2 substrate is used for our sensors. Oxygen plasma device found in the cleanroom is used for the bonding process. For the SiO_2 substrate, the process is as follows:

- (i) Clean SiO_2 wafer surface using acetone followed by isopropyl alcohol.
- (ii) Place both the PDMS channel and the wafer into the oxygen plasma chamber.

- (iii) Start the machine with following parameters:
 - (i) RF Power: 85 Watts
 - (ii) O₂: 25 sscm
 - (iii) N₂: 5 sscm
 - (iv) Chamber Temperature: 30° C
- (iv) Carefully place the channel over the sensing region using an alignment machinery.
- (v) Seal the channel using extra PDMS on the sides and at tubing ports.

This is the bonding process for PDMS and SiO_2 wafer. A similar approach could have been taken to bond PDMS and FR4 substrate. However, FR4 is unsuitable for oxygen plasma treatment. PDMS does not stick to the surface after treatment. Due to this, new versions of oscillator boards have been designed.

3.4.1 Microchannel Integration to FR4 Board and v5.x Series

Oxygen plasma works by cleaning organic materials with highly reactive oxygen plasma. This leaves silanol (SiOH) groups on the surface [34]. Following oxygen plasma treatment, the PDMS is placed in contact with the SiO_2 wafer, forming Si-O-Si bonds at the interface. These bonds are strong, and create a water-tight seal even under moderate pressure.

With FR4, these bonds do not form. Therefore, the surface of FR4 has to be covered with another, suitable material. However, this material should not act as an insulator between the flowing liquid and the resonator conductor. The surface is covered with solder resist satisfying both of these criteria. A slit on the solder mask has been opened, creating a channel for fluid flow. PDMS microfluidic channels can be bonded to this surface, as seen in Figure 3.8.b.

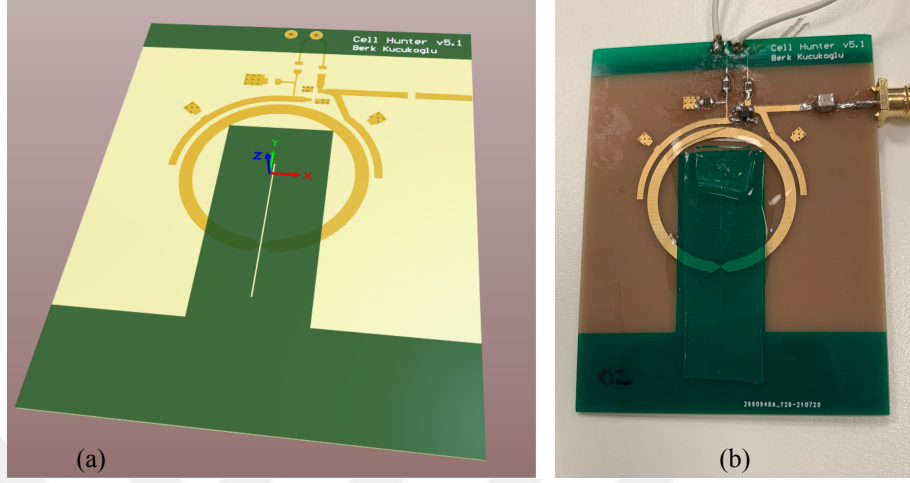


Figure 3.8: (a) Oscillator layout designed in Altium Designer, and (b) manufactured design with integrated microfluidic channel.

3.5 Preliminary Experimental Setup

DC voltage at 5V is fed to the oscillator from a linear voltage regulator, L7805. The output of the oscillator is connected to Rohde&Schwarz HM8123 Frequency Counter. After connecting the microfluidic tubings, the oscillator board is placed under the microscope as seen in Figure 3.9. A microscope is used to validate frequency shifts caused by the passing of microparticles visually.

In the first experiment, DI water without any particles is injected into the channel. This is done as a control experiment, measuring the Allan Deviation [35] under steady-state operation. Only a limited span of time-constants are available to us due to the limitations of the HM8123 instrument. This is explained in detail at Section 3.7.

In the second experiment, a mixture involving polystyrene microparticles with $30\mu\text{m}$ diameter is prepared and fed into the microfluidic channel. Once again, Fluigent MFCS is used as the pump. During the experiment, oscillation frequency values of the oscillator are recorded. Video showing events at the sensing region, as seen in Figure 3.10, is also recorded.

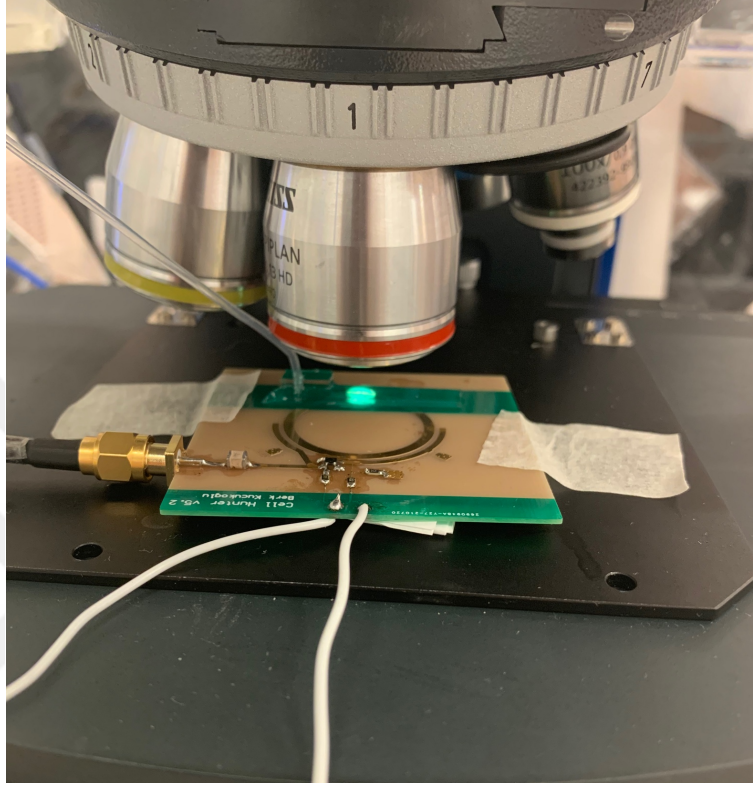


Figure 3.9: Oscillator board is placed under the microscope for the experiment.

3.6 Preliminary Experimental Results

Data for the first (Allan deviation) experiment can be seen in Figure 3.11. The plot shows that the Allan deviation is good for high-frequencies denoted towards the left side of the time-period axis. On the other hand, stability for longer time scales is poor. This trend can also be seen from the time domain data below.

Results for the $30.0\mu\text{m}$ polystyrene experiment can be seen in Figure 3.12. Data in the figure is cleaned with wavelet transform. Signal is decomposed with Daubechies 6 wavelet at 7 levels of decomposition. Parts with least amount of energy are thrown away.

Looking at the time domain plot, notice the significant variations in the mean at long time periods. This almost-sinusoidal time variation can be due to thermal fluctuations. The effects of this variation can also be seen back at the Allan

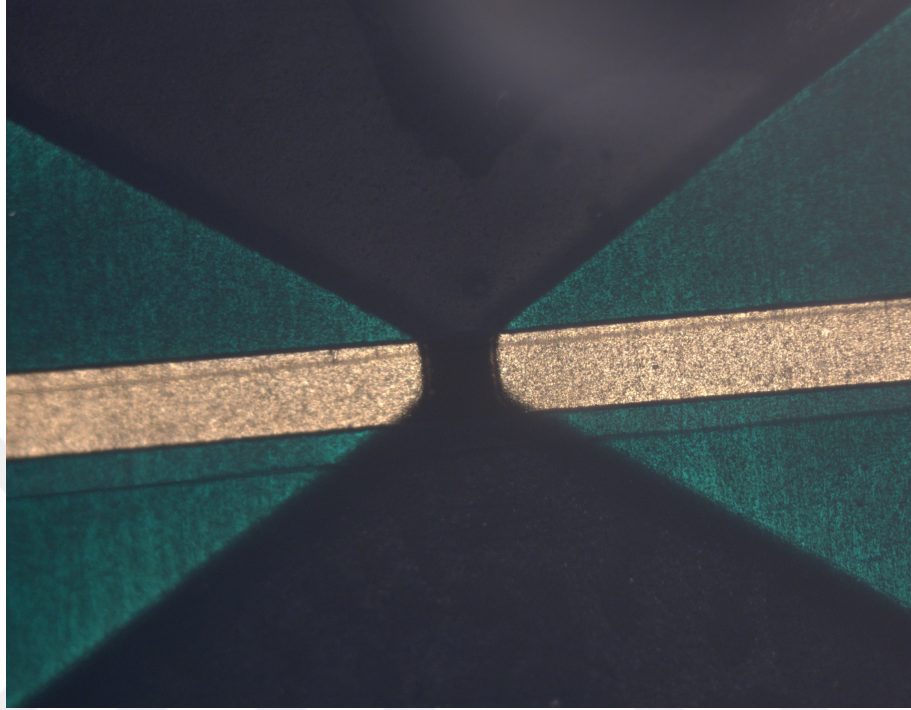


Figure 3.10: Snapshot from the recorded microscope video, focused at the sensing region.

deviation plot. Quality of Allan deviation decreases with increasing τ .

The “fast fluctuations” correspond to polystyrene passage events, validated from microscope images. Frequency shifts for events are not very significant. Polystyrene events shift the frequency around a couple of kHz, compared to CSRR where the mean amplitude of events was approximately 160kHz. There can be many reasons for this. For example, considering the sampling rate is five times slower than the CSRR setup, the peak forms may not be evident due to inadequate sampling rates. These are discussed in the next section.

3.7 Problems with the Setup

It is mentioned in the previous section that not a wide span of Allan deviation values are available to us. The reason behind this is the limited data rate due to

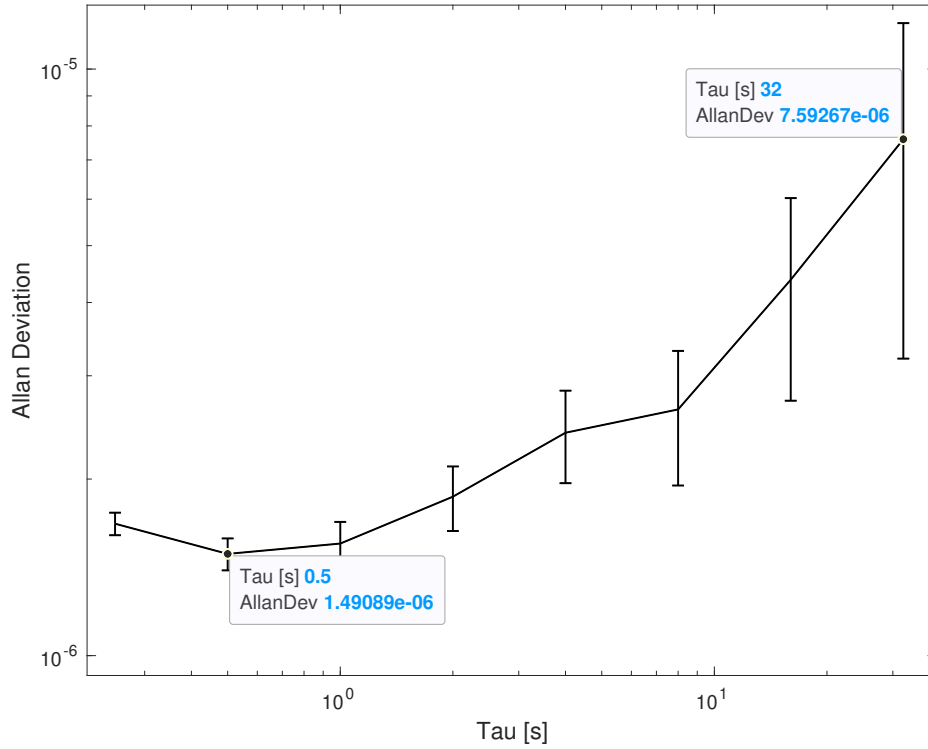


Figure 3.11: Allan Deviation plot for evaluating oscillator stability.

the HM8123 counter. The device takes more than 100ms to generate a frequency reading. This reading is then transferred to the PC running a MATLAB script. The rate of the asynchronous data transmission is fixed at a rate of 9600 bits/s. Furthermore, after a reading is performed, the counter requires some “dead-time” before being able to generate the next data point. Consequently, the data rate is limited to around 4 samples per second, or a 250ms period per sample. Therefore, stability data beyond this point is not available to us.

The data-rate limitation also created problems during the microparticle experiments. Passage of particles through the sensing region takes approximately 400ms to 1000ms, proportional to the flow speed. This means there is time to take roughly 2 to 4 data points per event, and that is extremely low to derive any information about the particle.

Our solution to the problem was to design a custom data processing board.

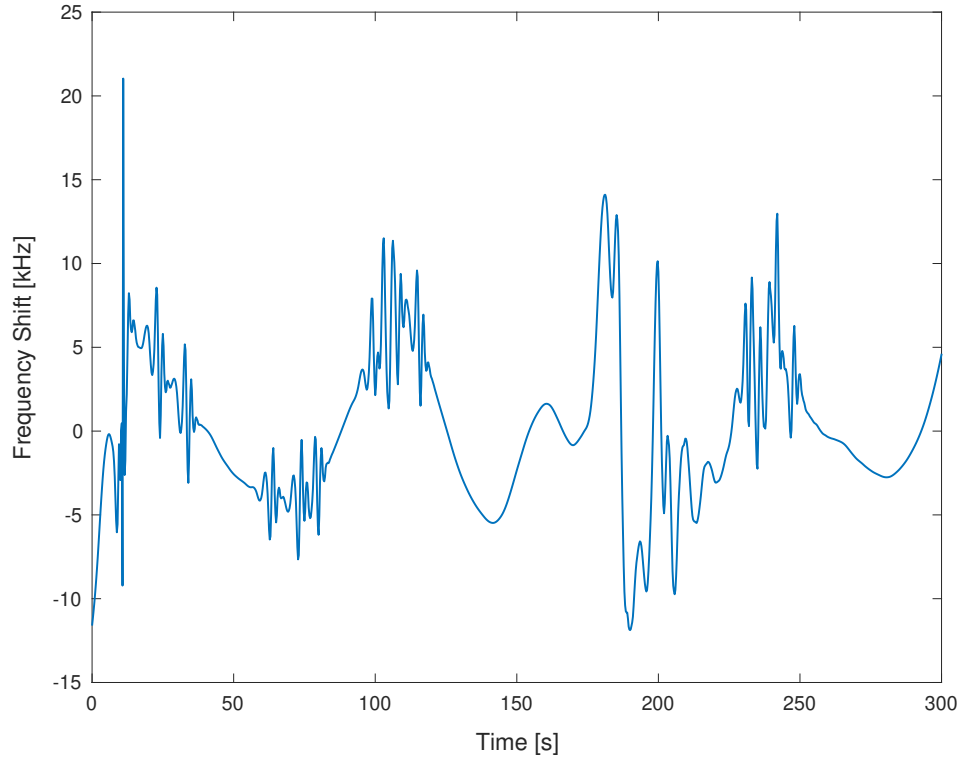


Figure 3.12: Time domain plot for the oscillation frequency during the polystyrene experiment (denoised with wavelet transform).

We suggest a microcontroller board for sampling the data with a high-throughput ADC, and transmitting this data to the PC for storing. Step by step, problems with this approach are identified and solutions are provided. All solutions are implemented, except for the last step where a plan is made for future studies.

The microcontroller board can be seen in Figure 3.13. Microcontroller chip dsPIC33CH256MP205 from Microchip company is used. It is a dual-core microcontroller capable of performing 100M multiply-and-accumulate (MAC) operations per second. It has an on-board ADC operating at a rate of 3.5M samples/s.

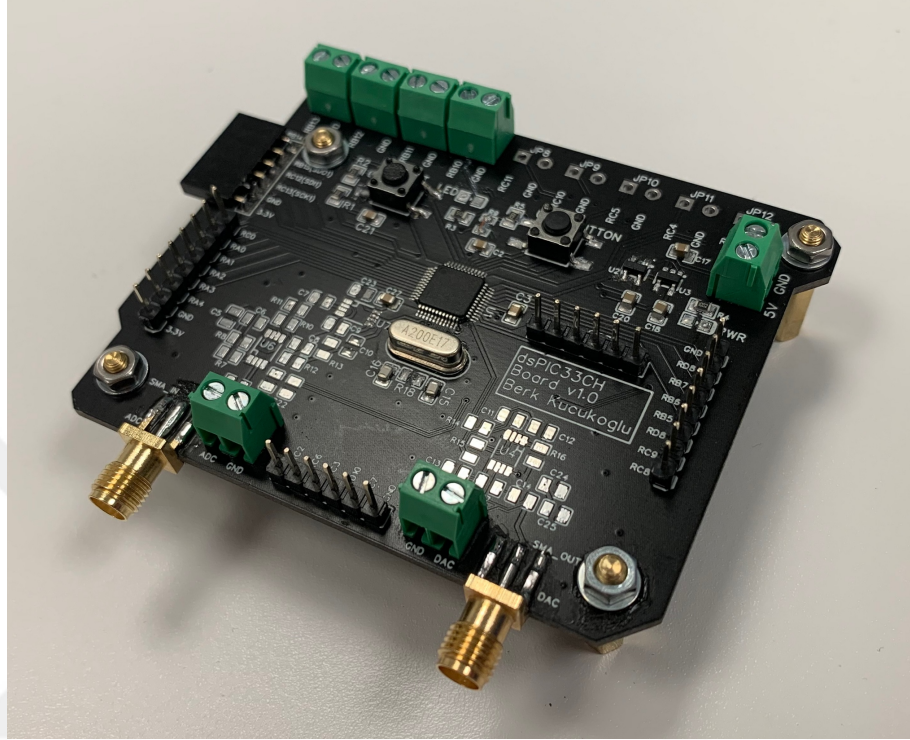


Figure 3.13: Designed board for real-time data processing. At the heart of the design lies the dsPIC33CH microcontroller.

3.8 Problem I: Down Conversion

The oscillator's operating frequency is around 2 GHz. Expected frequency shifts of events are around 100kHz. Therefore, with some safety factor, the required (positive) bandwidth is around 300kHz. To properly (in Nyquist sense) sample this, ADC has to operate at 600k samples/s, minimum. The plan, illustrated in Figure 3.14, is to sample the incoming data and transmit it to the PC for recording.



Figure 3.14: Illustration for the initial plan. Oscillator's output will be downconverted, sampled, and transmitted to PC.

To sample the oscillator's output, first it has to be downconverted. The downconversion process is done using a fixed frequency signal generator and a wideband mixer. The selected frequency for IF is zero Hertz, or DC. This is also named as homodyne detection [36]. The method in the frequency domain can be seen in Figure 3.15.

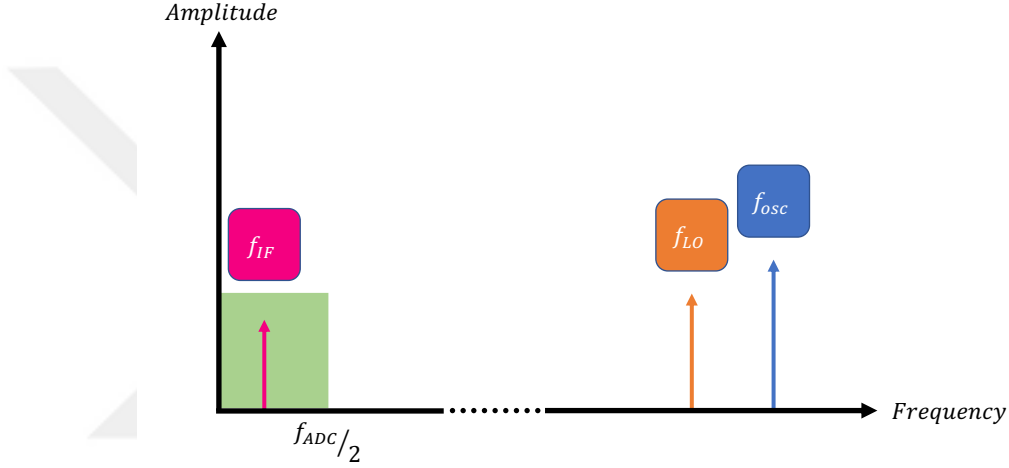


Figure 3.15: Illustration of the (naive) downconversion process. Green area is the bandwidth sampled by the ADC.

When we experimented with this method, it was observed that the downconverted signal was lost in the noise floor. At higher IF frequencies (e.g., 6 MHz), this problem is seen to vanish. It is inferred that the problem is due to the noise process called “flicker noise” or “pink noise” [37]. The power spectrum of this noise process increases with lower frequencies. It is extremely limiting for our case, which is at a bandwidth from DC to 300kHz.

3.8.1 Solution I: Non-Zero IF for Down Conversion

As the problem is identified in the previous section, the solution is to use a higher intermediate frequency for the downconversion process. As a starting point, a downconversion frequency of 6.15 MHz is chosen.

With this new plan, undersampling has to be performed. The term is explained

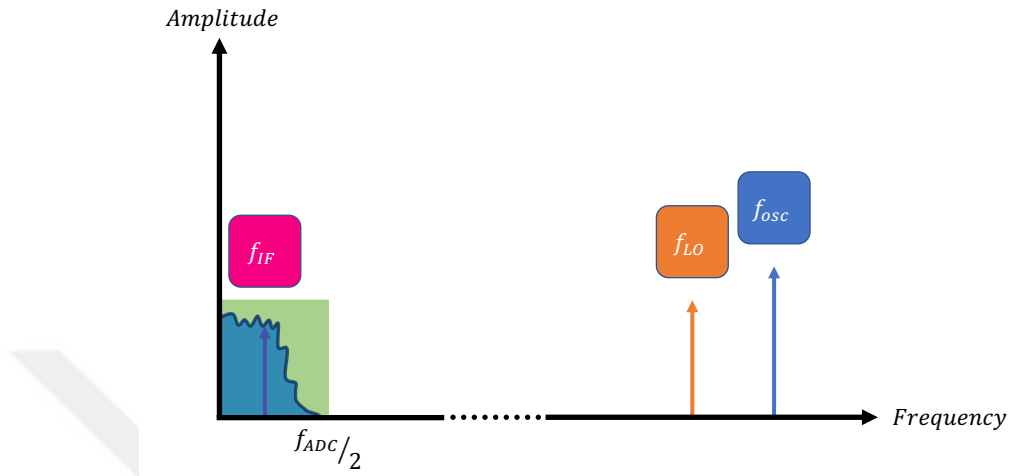


Figure 3.16: Illustration of the problem. Downconverted peak is lost in the Flicker Noise process.



Figure 3.17: Illustration for the improved plan. ADC will perform undersampling of the signal.

in detail in [38]. Therefore, the sampling frequency must be chosen carefully. For a sampling frequency of 600kHz, the aliased component will sit at 0.15 MHz. There is 150kHz of room to move up or down, before getting negatively affected by aliasing.

The solution is implemented in both hardware and software. From a theoretical standpoint, the illustration of the undersampling process to avoid flicker noise can be found in Figure 3.18.

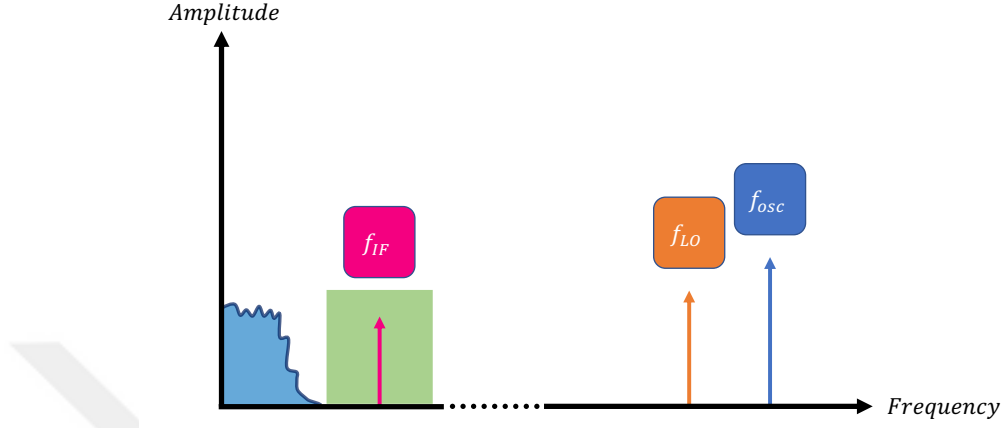


Figure 3.18: Explanation of the method to avoid flicker noise process during sampling.

3.9 Problem II: High Data Rates

With the last improvement, the ADC speed is fixed to 600k samples/s, where each sample is 12-bits long. All of these samples have to be transmitted to the PC and stored. This requires a data transmission rate of:

$$12 \frac{\text{bits}}{\text{sample}} \times 600k \frac{\text{sample}}{\text{s}} = 7.2 \text{Mbits/s.} \quad (3.3)$$

This rate is unattainable with UART communication. Furthermore, the situation is even worse because UART communication uses 8-bit long packages. Therefore, one sample (12-bits) requires two transmissions. In addition, sending each package requires one start and one stop bit to be transmitted. Therefore, another method is required.

3.9.1 Solution II: Real-Time Signal Processing

The solution to the problem lies in real-time signal processing. We are trying to measure the oscillation frequency of our oscillator. The signal we are interested in is a sinusoid between DC to 300kHz. When Fourier transform is applied to this signal, a peak will be observed in the frequency domain. We are interested

in the location of this peak point.

It is clear that we do not have to transmit all of the data. By processing data with the microcontroller board in real-time, only the location of the peak can be transmitted.

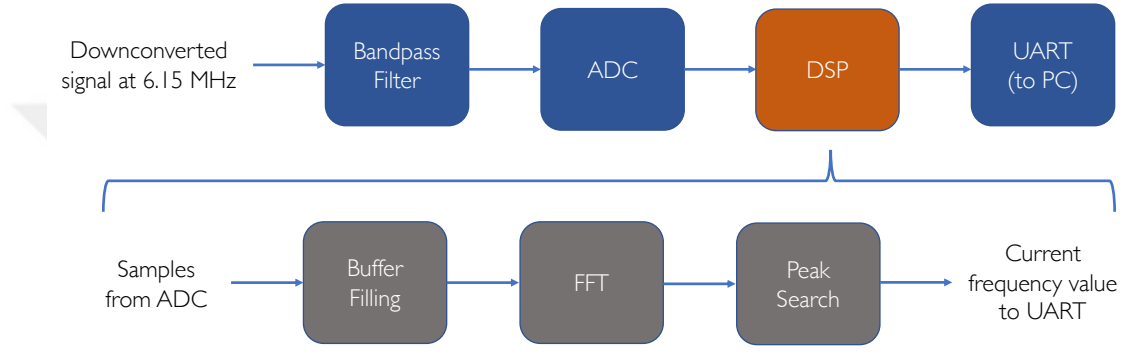


Figure 3.19: Illustration for the improved plan, second iteration. Fast Fourier Transform will be evaluated on the run. Only the peak location will be transmitted.

With this method, a data-transmission rate reduction on the order of thousands can be achieved. Details of software implementation are covered in the following subsection.

3.9.2 Software Design and Code Optimization with Assembler

A long, empty array is created for buffering purposes. The length of the buffer can vary; however, a length of 1024 samples can be a good starting point. ADC is configured to sample the analog signal continuously. Digital samples are loaded into the buffer one by one as they arrive. The code snippet can be found in Section B.1.1. When the buffer is full, an interrupt is generated. In this interrupt routine, Discrete Fourier Transform is evaluated with the Fast Fourier Transform (FFT) algorithm. After the evaluation, the magnitude of the resulting complex array is calculated, and the peak location is found. Later, the location is sent to the UART module to be transmitted to the PC.

With this scheme, only a single value is sent every time the buffer is full. Therefore, assuming the buffer is 1024 samples long, the required data transfer rate is lessened by three orders of magnitude. There is a direct proportion between the buffer length and the data transfer advantage.

Microchip products, such as the dsPIC used in this project, can be coded in the MPLABX programming environment. The compiler provided by Microchip is XC16, a special compiler for Microchip Microcontrollers based on the C programming language. Using C language for the code usually gives a satisfactory performance. However, for very high-performance tasks, functions written in C language do not yield adequate results. As an example, when the buffer is full, the FFT is calculated. Even though the dsPIC devices include built-in hardware required for single-cycle multiply-and-accumulate operations, code that is written in C language does not make use of this hardware support. In addition, to operate on an array with loops in C wastes a significant number of instruction cycles, due to overhead and other complications.

The code written in C takes more than 30,000 cycles for an array of 1024 samples. On the other hand, the loop in Assembly takes around 1055 cycles for the same array. The difference between the two is that the C code wastes many instructions during the for-loop. The Assembly version produces a “no-overhead” loop with the REPEAT function. For our case, writing time-critical parts in Assembly has reduced the FFT calculation time to approximately 0.83ms for an array of 1024.

3.10 Problem III: Unstable Oscillation Frequency

After all of the solutions are implemented, an experiment is performed to see the results. Unfortunately, the oscillator frequency is not stable in the 300kHz bandwidth. This can be due to many reasons. The most significant for poor

stability is the thermal drift. As seen in the Allan Deviation plot at (Figure 3.11), the oscillator gives good performance at high frequencies. However, the thermal drift is significant.

Our ADC is set to read a bandwidth of 0-300kHz. As thermal drift begins taking effect, the oscillation frequency changes drastically. After a short duration, the oscillation frequency floods the bandwidth. This causes aliasing effects, and the oscillation frequency cannot be appropriately measured. A method has to be implemented to eliminate low-frequency drift and keep the oscillator in the ADC bandwidth.

3.10.1 Potential Solution III: Oscillator Board V5.4

The latest version of oscillator boards, v5.x series introduced in Section 3.4.1, is designed to experiment with new concepts. For the solution of problem III, we will focus on oscillator board v5.4, as seen in Figure 3.18.

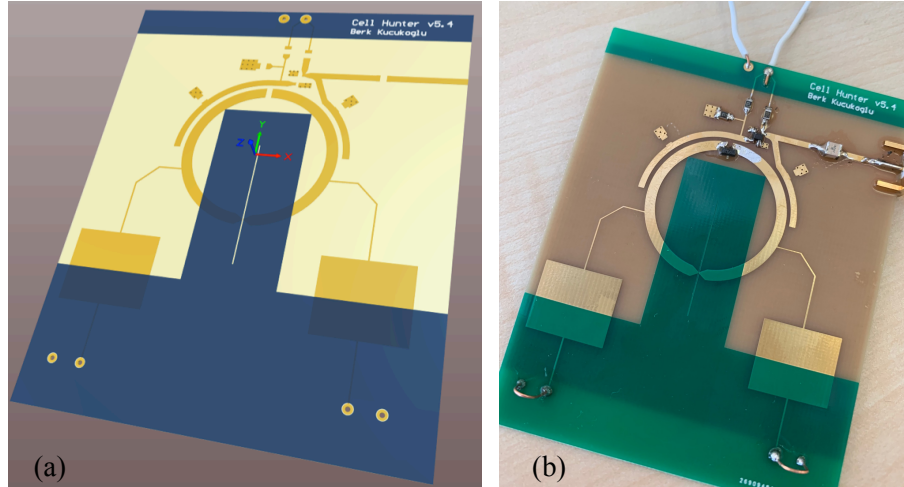


Figure 3.20: (a) Oscillator layout designed in Altium Designer, and (b) manufactured design. Notice the second (larger) gap at the top part of the resonator.

In this oscillator version, a second gap is introduced at the top part of the resonator. This gap is used for soldering a varactor diode onto the resonator. The frequency of oscillations can be controlled by using this diode. Varactor

diodes have a variable capacitance, and by applying a DC voltage, the oscillation frequency can be changed. After integration with a control loop, this phenomenon can be used to eliminate the drift caused by temperature or other fluctuations during experimentation.

The varactor diode used is SMV1275 diode from Skyworks company. It is a high-frequency varactor diode suitable for microwave applications. The capacitance can change from 4.74 pF (at 1V) to 0.89 pF (at 12V). We plan to use the diode with our dsPIC controller board. It has an integrated on-chip DAC, with a range of 0-3.3V at a resolution of 16-bits.

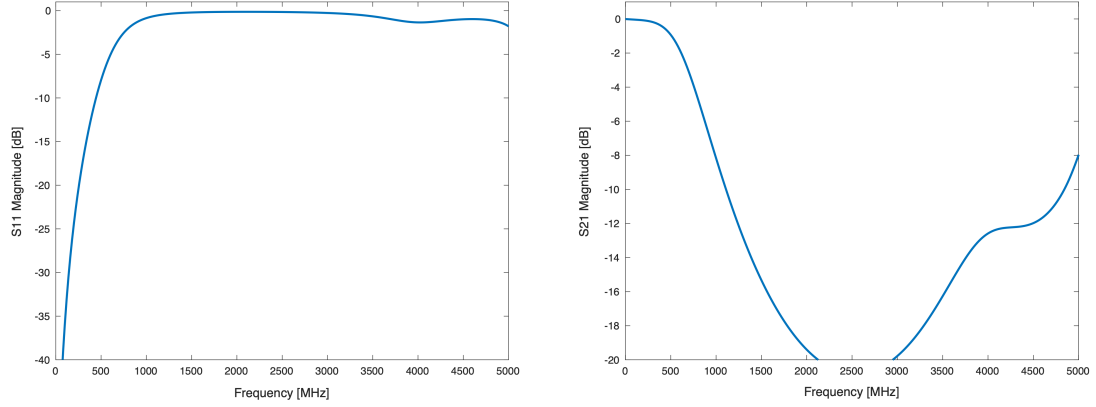


Figure 3.21: S11 and S21 plots for the isolator.

There are two large conductor pads on each side of the resonator. These are distributed element lowpass filters in order to isolate the DC feed points from the oscillator. This way, the DC feed points are invisible to the resonator, whereas the user can feed DC to the varactor diode simultaneously. In mathematical terms, the filter structure has a very flat S11 response at the oscillation frequencies as seen in Figure 3.21. Signals sustained on the resonator will not be able to pass to the DC feed points. On the other hand, the S21 (and S12) plot is flat around DC frequencies, meaning the user can easily feed DC voltage to the circuit.

The success of the design owes to the optimization process performed in ADS Keysight simulation software. After the filter type is determined, which is the Stepped-Impedance transmission line geometry, a microstrip transmission line consisting of 6 pieces of smaller transmission lines with variable width/length are

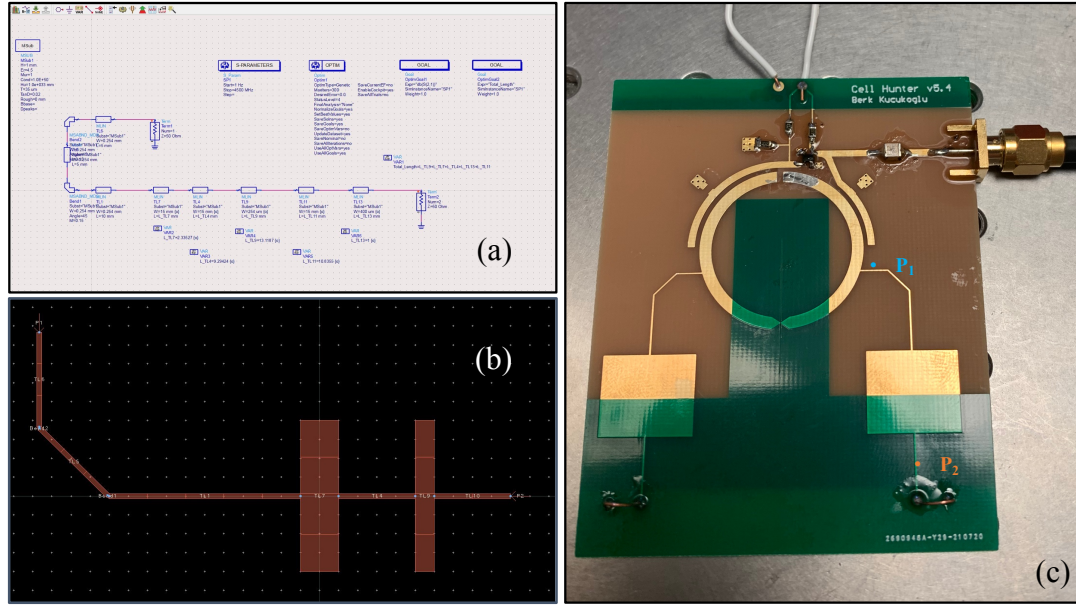


Figure 3.22: (a) Schematic view of the optimization window. (b) Layout view of optimization view. (c) Manufactured design of the oscillator, before diode is soldered.

placed into the simulation software. Then an optimal design is reached using a basic form of genetic algorithm. The optimality cost function is calculated in the sense of flattest (and highest) S_{11} response in oscillation frequency. Perfect transmission at DC is already guaranteed due to the nature of the stepped-impedance filter structure.

With oscillator board v5.4, the frequency of the oscillations can always be contained in the ADC sampling range. Significant fluctuations can be eliminated using a PLL-like approach. The solution will be implemented in the near future, as discussed in the conclusion chapter.

Chapter 4

Conclusion and Future Plans

In this chapter, a summary of the thesis will be given. Future plans for these projects will be visualized with an emphasis on potential solutions.

4.1 CSRR Sensors and Integration with Other Sensors

Concentric split-ring resonators are used to detect single cells and microparticles while also distinguishing the differences in between due to the electrical volume. Dielectric permittivity is the main factor creating this distinction. This idea is proven with experiments. Results show CSRR is a much better geometry compared to the older generation of Bowtie devices, giving at least an order of magnitude better response.

The most critical issue related to this sensor is the problem of positional sensitivity. Cells and microparticles do not always pass directly from the sensing region. The sensitivity of the sensor drops very rapidly away from the gap. Slight deviations in the pathway of microparticles result in a significant difference in the generated electrical signal.

A similar problem is the z-axis dependency of the sensor. Sensitivity also drops down with the channel height. Experiments done with a channel height of $50\mu\text{m}$ give approximately double the response compared to experiments done with $100\mu\text{m}$ channel height.

A solution to these problems can lie in the design of a more complex microchannel structure. For example, the microfluidic channel can taper at the location of the gap, eliminating any deviation from the sensing region. A similar solution can also be done for the z-axis dependency. Channel can taper in 3-dimensions to eliminate all kinds of positional dependency. Another solution can be by using a different sensor geometry. The CSRR sensor is deposited on a wafer and is 2-dimensional in geometry. A more creative design with 3-dimensional geometry can be constructed.

4.2 Standalone SRR Integrated Oscillators

Oscillators integrated with split-ring resonators are used to detect polystyrene particles. From an EM perspective, they are using very similar resonator topologies. The difference lies in the measurement setup used to read from these sensors.

Problems concerning the CSRR sensor setup were the system's slowness, the parameter dependency of the results, and the cost-effectiveness of the measurement setup. With SRR integrated oscillators, these problems can be solved.

The system is much faster in operation because there are no delays due to communication between instruments or control loops working on a PC environment. With a sampling rate of 600k samples/s and a buffer length of 1024 samples, the calculation of the frequency value takes less than 2ms. Compared to the CSRR measurement system, where every frequency value arrives at 50ms intervals, this is a 25 times improvement.

For the PLL system, both K_p and K_i parameters are set before each experiment. This “setting” period usually consists of a trial-error process. Setting the correct parameters can result in an excellent signal-to-noise ratio. However, failure to set these parameters can also claim the sensor completely insensitive. With the oscillator setup, no control loops track the sensor’s resonant frequency to keep the system at resonance. Instead, the system is always at resonance. As particles pass, the resonator returns to its original frequency at its own pace and changes the output frequency. Therefore, there is no parameter dependency in the results. Each experiment is guaranteed to work the same way.

The cost-effectiveness is improved by eliminating instruments such as lock-in amplifiers, high-frequency variable signal generators, mixers, and circulators. The only requirement for the oscillator setup is a signal generator working at a fixed frequency and a mixer. These can even be integrated onto the oscillator board with newer versions.

The only valid problem at the moment is the unstable oscillation frequency. The short time stability is good. However, long-time stability is poor because of slow effects such as temperature fluctuations. These slow effects can be eliminated through the usage of a slow PLL-like control loop. This should not be confused with the control loop driving the resonator at the CSRR setup. This is more like a long-time stability solution to ensure the signal being sampled stays in the proper undersampling boundaries. This solution will be implemented in the near future.

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Appendix A

Supplementary Material for CSSR Sensor

A.1 COMSOL Simulation Mesh Settings

Four different mesh settings are used for three different domains on the geometry. These are the SMA connector domain, air volume domain, and substrate domain. The mesh settings are standard COMSOL settings of type “Normal” and “Physics-Controlled” for the SMA connector and air volume domains.

For the substrate domain, there are two types of meshes used. In the vicinity of the sensing region, a custom mesh setting is used with settings below:

- Maximum element size: 0.04 mm
- Minimum element size: 0.0005 mm
- Rest of the settings: From sequence (extremely fine)

This creates a highly dense mesh structure in a very localized area: the sensing gap. Typically, COMSOL software attempts to mesh the structure with

a “physics-controlled” approach. This means an element size on the order of 1/20th of the wavelength of interest is adequate for the simulation. However, for visualization of the electric field in the $20.0\mu\text{m}$ gap, many mesh elements have to be squeezed into the area. A less detailed mesh setting is used for the rest of the substrate: the “Finer” standard-setting.



Appendix B

Supplementary Material for Standalone SRR Oscillator

B.1 dsPIC Control Board Codes

B.1.1 Code Snippet: Filling the Buffer

```
int bufferADC[1024];
unsigned int SampleIndex = 0;

void NewSample_Interrupt()
{
    bufferADC[SampleIndex] = newSample;
    SampleIndex++;
}
```

B.1.2 Code Snippet: Calculating the Frequency Bin

```
int bufferADC[1024];
```

```
int BufferFullInterrupt()  
{  
    FFT_Array = computeFFT(bufferADC);  
    Mag_FFT_Array = computeMag(FFT_Array);  
    peak_Location = searchPeak(Mag_FFT_Array);  
    UART_Send(peak_Location);  
  
    return success;  
}
```

Appendix C

Detailed Design Dimensions for Microwave Resonators

C.1 Dimensions for CSRR

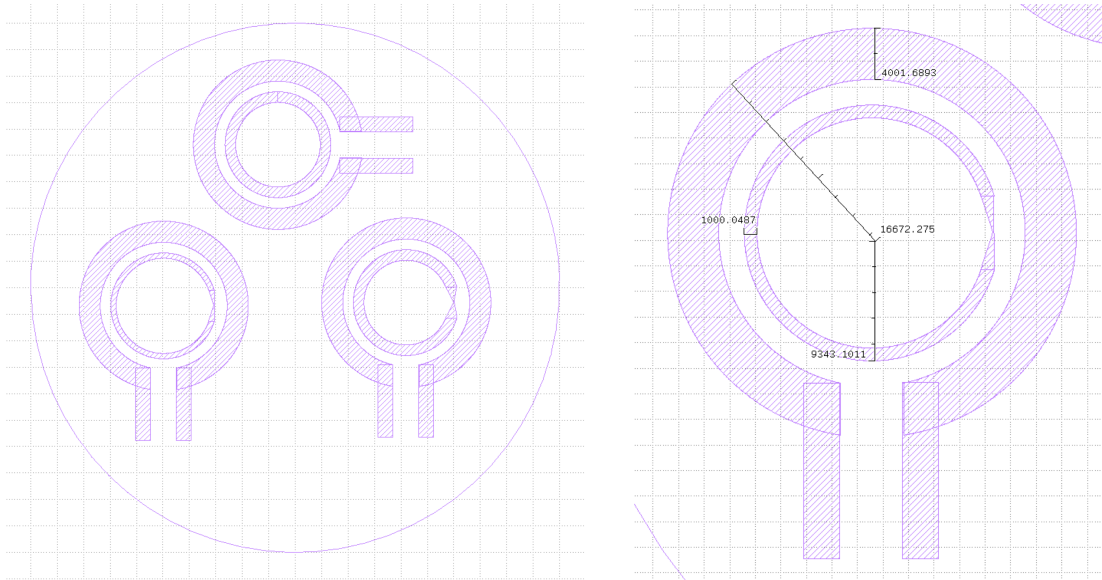


Figure C.1: Dimensions for the CSRR sensor. All dimensions are in micrometers.

C.2 Dimensions for SRR Integrated Oscillator

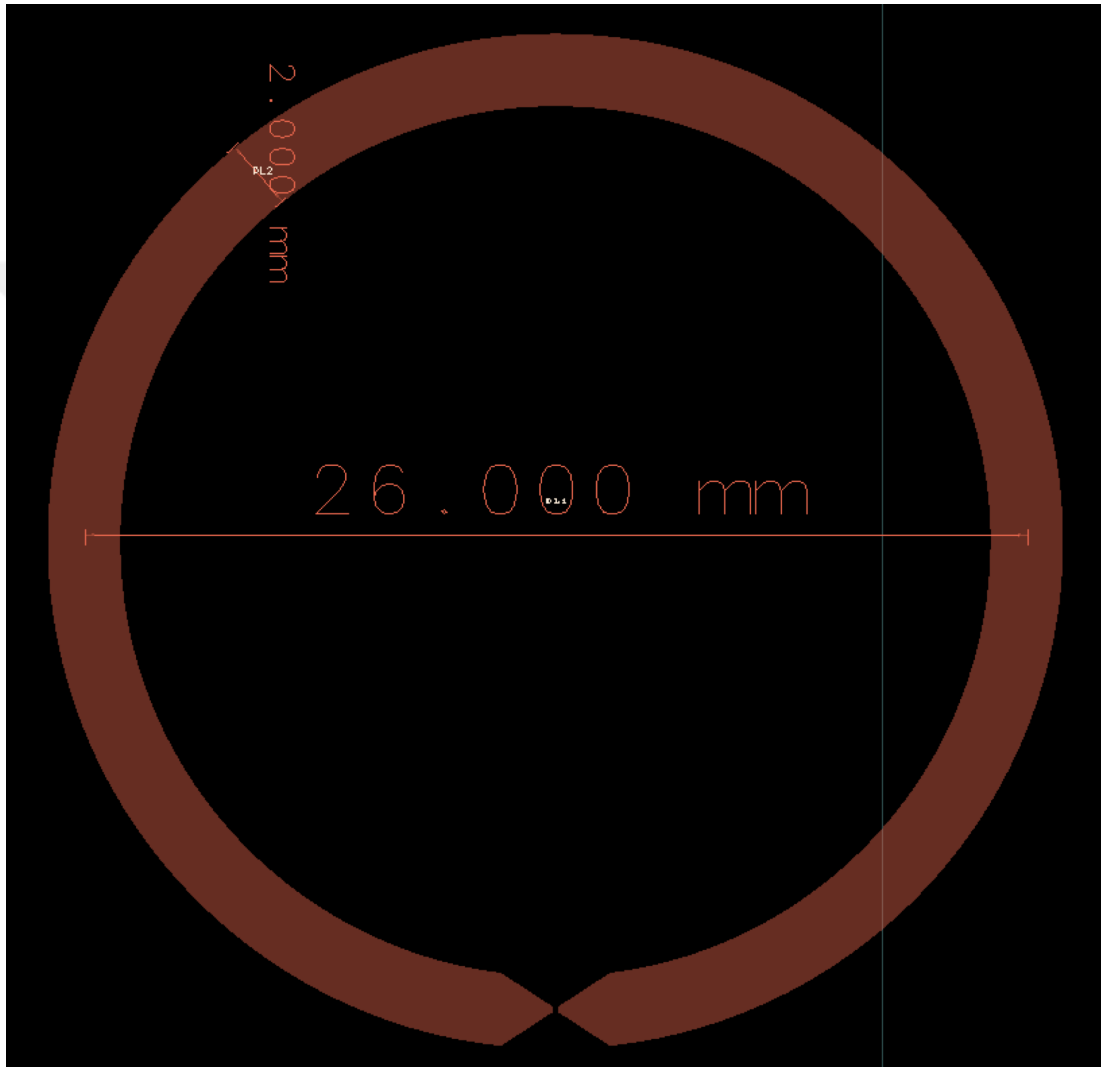


Figure C.2: Dimensions for the resonator section of the SRR integrated oscillator.

Appendix D

User Guide for dsPIC Data Processing Setup

In this appendix chapter, a simple user guide is presented. Software and hardware requirements are listed. Codes are provided to readers for repeatability.

D.1 Hardware Requirements

There are two major hardware requirements for the operation of our system: specifications of the microcontroller and the PC.

The microcontroller must be one of the dsPIC model microcontrollers from the Microchip company. For our case, this is the dsPIC33CH256MP205. For a cost-effective solution, the microcontroller can also be chosen as dsPIC33FJ128MC802.

The PC for running the MATLAB code must be capable of supporting the R2021A version of the MATLAB environment. Required hardware requirements, such as minimum RAM, can be found on the official MATLAB website. The PC should also have a spare USB port for UART-USB communications.

For programming the dsPIC microcontrollers, reader has many choices. In this thesis, the PICKIT3 programmer is chosen for the hardware. For USB-UART communications, any USB-to-UART communication chip can be used. We are using the Adafruit USB-to-UART converter board.

D.2 Software Requirements

The required operating system for the PC is Windows 10. Older versions of this operating system might also be adequate but may require minor changes in the steps provided in the next chapter.

For programming the dsPIC microcontroller, the MPLABX IDE environment is needed. The environment version should be no less than V5.50, considering the support for dsPIC33CH microcontrollers has been introduced with recent packages.

For USB-UART communications with the dsPIC device, the MATLAB programming environment is chosen. Other software can also be adequate but may require additional effort on the reader's side.

D.3 Steps for Proper Operation with dsPIC

1. Start the MATLAB program.
2. Set the required rate for communication. For our case, it is set as 115200 bits/s.
3. Connect dsPIC board to the PC through Microchip validated programming hardware (ex: PICKIT3)
4. Start MPLABX software.
5. Program the dsPIC microcontroller with the provided code.

6. Unplug the programmer.
7. Connect the USB-UART communication tool.
8. Connect the downconverted signal to the ADC port of dsPIC board.
9. Check if the dsPIC sends the value of the current frequency to the MATLAB environment.
10. Set the duration of the experiment. For our case, the duration is usually set to 10 minutes.
11. Start the MATLAB data logging program with the mentioned settings.
12. Plot the data through MATLAB built-in functions to observe the frequency-time graph.
13. Finalize the experiment.

D.3.1 dsPIC Code

Listing D.1: dsPIC Code

```
/*  
  
* File: ThesisBoard_v1.c  
  
* Author: Berk Kucukoglu  
  
*  
  
* This code is for my thesis.  
  
* A buffer of FFT_BLOCK_LENGTH samples is created. This buffer is filled  
  
* an ADC working at a given speed.
```

** The time this buffer is filled , the microcontroller performs an FFT.*

** Result of this FFT is transmitted to the PC over UART working at a*

** rate of 115200.*

**
This device can work upto 40 MHz, in terms of Fcy.
Fcy is equal to Fosc/2
Hence we aim for Fosc of 80MHz*

** Created on September 06, 2021*

**/*

#include <xc.h>

#include <stdlib.h>

#include <dsp.h>

#include "configuration_settings.h"

#include "functionList.h"

```

//#include "testData_Berk.h"

/* .....

// Definitions

#define FFT_LENGTH 1024 // Number of frequency points in the FFT

#define LOG2_FFT_LENGTH 10 // Number of "Butterfly" Stages in FFT

// FFT Related

void FFT_Setup(void);

void FFT_Run(void);

fractional FFToutput[FFT_LENGTH/2];

int16_t peakFrequencyBin = 0; // Declare post-FFT variables to compute

uint32_t peakFrequency = 0; // frequency of the largest spectral compon

fractcomplex twiddleFactors[FFT_LENGTH/2]

__attribute__((section(".xbss,bss,xmemory"),aligned (FFT_LENGTH*2)));

fractcomplex sigCmpx[FFT_LENGTH]

__attribute__((eds, space(ymemory), aligned (FFT_LENGTH * 2 * 2)));

```

```
//Global Variables
```

```
unsigned int debug_checker = 0;
```

```
int twidGen = -1;
```

```
unsigned int globalIndex = 0;
```

```
unsigned int FFT_Run_Request = 0;
```

```
/* .....
```

```
// Interrupt Routines
```

```
void __attribute__((interrupt , no_auto_psv)) _T3Interrupt(void)
```

```
{
```

```
    IFS0bits.T3IF = 0; // Clear Timer3 interrupt flag
```

```
}
```

```
/* .....
```

```
void __attribute__((interrupt , no_auto_psv)) _ADC1Interrupt(void)
```

```
{
```

```

sigCmpx[globalIndex].real = ADC1BUF0;

sigCmpx[globalIndex].imag = 0;


if(globalIndex<FFTLLENGTH-1)
{
    globalIndex = globalIndex + 1;
}
else
{

    AD1CON1bits.ADON = 0;    // Turn off the A/D converter

    globalIndex = 0;

    FFT_Run_Request = 1;

}

IFS0bits.AD1IF = 0; // Clear ADC1 interrupt flag

}

/*.....

```

```
// Main Code
```

```
int main(void)
```

```
{
```

```
// Pin Configuration
```

```
configure_Pins(); // This is important because the pins are con
```

```
// according to the hardware board that the uC is built upon.
```

```
// Clock Configuration
```

```
configure_SysClock();
```

```
// Don't touch above this line
```

```
// Variables
```

```
//
```

```
// Peripheral Settings
```

```
configure_Interrupts();
```



```

configure_Timer3 ();

configure_ADC ();

configure_UART ();

configure_DSP ();

FFT_Setup ();

// Program Begin

while (1)
{

    if (FFT_Run_Request == 1)

    {

        FFT_Run ();

        U1TXREG = (peakFrequencyBin & 0b0000000011111111);

        U1TXREG = (peakFrequencyBin & 0b1111111100000000);

        FFT_Run_Request = 0;

        AD1CON1bits.ADON = 1;    // Turn on the A/D converter

    }

```



```

void FFT_Run()

{

    // Remember, the FFT function requires input data to be in the
    // fixed-point range,  $[-0.5, +0.5]$ 

    FFTComplexIP(LOG2_FFT_LENGTH,&sigCmpx[0],&twiddleFactors[0],COI);

    BitReverseComplex (LOG2_FFT_LENGTH,&sigCmpx[0]);

    SquareMagnitudeCplx(FFT_LENGTH/2,&sigCmpx[0],FFToutput);

    FFToutput[0] = 0; // Eliminate DC

    FFToutput[1] = 0;

    VectorMax(FFT_LENGTH/2, FFToutput, &peakFrequencyBin);

}

```

D.3.2 MATLAB Code for Data Logging

Listing D.2: Matlab Code

```

%% General Settings

Fs = 100000;
bufferSize = 1024;

```

```

num_RequestedSamples = 5000;

dsPIC_SampPeriod = 517600; % Cycles
dsPIC_ClockSpeed = 40e6; % Cycles/second

%% Communication Settings

portName = "COM4";
baudRate = 115200;

%% Rest is Automatic
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%

Ts = 1./(Fs); % seconds
BintoFreqCoef = Fs./bufferSize;
TimeBetweenSamples = dsPIC_SampPeriod./dsPIC_ClockSpeed; % seconds

%% Create Object

externalUART = serialport(portName,baudRate,"Timeout",3);

%% Set Object Properties

externalUART.ByteOrder = "little-endian";
flush(externalUART);

%% Get Data

flush(externalUART);
data_FreqBins = read(externalUART,num_RequestedSamples,"uint16");

%% Create Plotting Arrays

```

```
n_index = 0:1:length(data_FreqBins)-1;  
time_index = n_index*TimeBetweenSamples;  
data_FreqVals = BintoFreqCoef*data_FreqBins;
```

```
%% Plot
```

```
figure  
plot(time_index , data_FreqVals*1e-3);  
xlabel( 'Time [s] ' );  
ylabel( 'Frequency [kHz] ' );  
axis tight
```