ÉCOLE POLYTECHNIQUE
FÉDÉRALE DE LAUSANNE

PCB Design and Fabrication for NEMS Applications

Advanced NEMS group

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Table

1. Introduction

Microelectromechanical systems (MEMS) are miniaturized devices that combine the electrical effects with the mechanical ones. The aim is to realize some sensors and actuators at a small scale. Downscaling some devices offers some great physical properties. MEMS can be found in cares, planes, biology, medicine, telecommunication etc. They emerge at the beginning of the 70's and the first commercialized MEMS were made at the beginning of the 80's.

Nanoelectromechanical sensors (NEMS) are the next logical step after MEMS. The downscaling of some devices close to or below the micrometers offers some new possibilities in terms of responsivity and resolution. Some are developed for telecommunication.

The printed circuit board (PCB) is a non-conductive mechanical support with some conductive tracks which connect different electrical devices like chips, surface mounted devices (SMD), resistor, capacitor etc. They can have many layers to have a better density of components. The layers can be connected by vias. Most of the mechanical support is made of FR-4 glass epoxy. The tracks are made of copper and are covered by a thin layer of isolation. We can find printed circuit boards everywhere, phones, computers, cars, the kitchen etc. It is the simplest way to drive a chip.

For this semester project a PCB has to be designed to drive a chip. A source voltage is put at the input of the chip. The main part of the chip is a flexural beam resonator and the aim is to maximize the power transferred to the detector which is put at the output of the chip.

First, we are interested in the theory. We talk about the flexural beam resonator and its functioning. The resonator has an equivalent circuit model which is represented by a RLC circuit with a capacitance in parallel called background noise. The concept of reflective waves will be introduced to understand how to drive the chip. To avoid reflective waves, we need to put some components between the chip and the source/detector. It is called the matching. The concept of two different matching is introduced. The matching with an inductor and capacitor (LC-Matching) and the matching with a transformer (Transformer-Matching).

Then, we simulate the electrical behavior of the chip. The results and the simulation of many designs are done. The values of the RLC resonator and the value of the background noise capacitance are computed. Quite Universal Circuit Simulator (QUCS) is used to analyze and maximize the transmitted S-parameter of all the different circuits. First, the resonator is simulated without the background noise capacitance to see its resonance frequency. A match of the resonator is made. Then the background noise capacitance adds complexity to the matching. We try to cancel this effect by adding some components in parallel of the chip. Several designs are tested in order to have the best peak of resonance frequency. The LC-matching and the Transformer-matching are compared. Two designs to cancel the background noise capacitance are also compared.

Based on the commercial availability of the components and the S-parameter results, one of these designs is chosen. The schematic library and the PCB library of all the components are made on Altium and a first design of the PCB is presented in this report.

2. **Theory**

2.1 Flexural beam resonator

The main part of the chip is a flexural beam resonator. The scheme is shown in Figure 1.

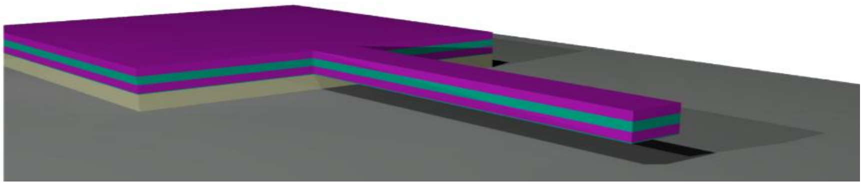

Figure 1 - Image of the flexural beam resonator

The flexural beam is made of a piezoelectric transducer (grey volume between the violet parts). An alternative (AC) excitation voltage V_{IN} is put on the electrodes (violet) to actuate the transducer with the direct piezoelectric effect. Then, the inverse piezoelectric effect is used by the same transducer to sense the amplitude and the frequency of the vibration of the beam. It gives the out voltage V_{OUT}. According to the dimension of the electrodes and the piezoelectric properties, the beam has a frequency response as shown in Figure 2. f_r is the resonance frequency of the device in which the amplitude of the vibration is maximum. Consequently, the spectrum of the frequency response influences on the out voltage. This is the behavior of a resonator.

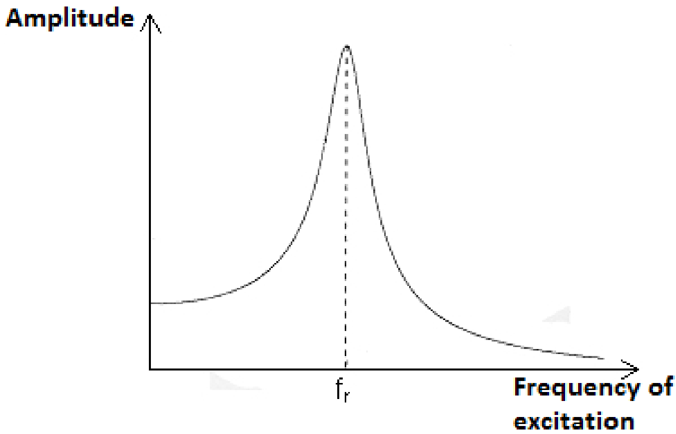

Figure 2 - Frequency Response

2.2 Equivalent Circuit model

The resonator can be represented by an equivalent electrical circuit. The equivalent circuit model of a resonator is defined by a motional resistor, capacitor and inductor (R_m, C_m and L_m) represented in Figure 3. R_m, C_m and L_m are defined by the flexural beam properties and piezoelectric effect. C_0 is called the “background noise” which is the current passing through the capacitor created between the two electrodes of the flexural beam (see Figure 1) since the thickness of the piezoelectric layer is small.

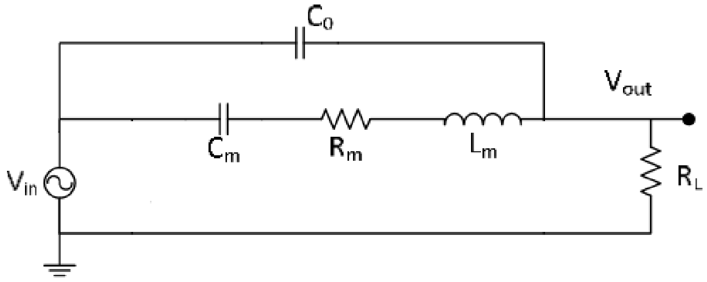

Figure 3 - Equivalent circuit of the resonator

To define the values of the impedances, the direct/inverse piezoelectric effect and the resonance frequency of the beam is used. The strain is defined by:

$$\varepsilon_i = C_{ij} \cdot \sigma_j + d_{ij} \cdot E_j$$

The displacement field is defined by:

$$D_i = \zeta_{ij} \cdot E_j + d_{ij} \cdot Y_{ij} \cdot \varepsilon_i$$

ε_i is the strain, C_{ij} is the compliance matrix, σ_j is the stress, d_{ij} is the piezoelectric constant, E_j is the electric field, D_i is the displacement field, ζ_{ij} is the dielectric constant and Y_{ij} is the young modulus.

By inserting ε_i in the displacement field equation and adding the resonating effect of the beam, the current density is then found by taking the time derivative of the displacement field. Then, the current density is integrated over the surface area to get the current:

$$I = \int_A \frac{\partial D_3}{\partial t} dA \approx j\omega C_0 V_{in} + 7.35 j\omega \cdot \frac{d_{13}^2 z_{offset}^2 LYb}{t_{total}^3} \cdot \frac{V_{in}}{1 - \left(\frac{\omega}{\omega_0}\right)^2 + j\frac{\omega}{\omega_0 Q}}$$

C_0 is the background noise capacitance, z_{offset} is the thickness centre of the PZE minus the total thickness centre of the beam, L is the length of the beam, b is the width of the beam, t is the total thickness of the beam, ω_0 is the resonance frequency of the beam and Q is the Q-factor of the resonator.

We can note that this equation characterizes two elements in parallel: The background noise passing through the capacitor C_0 and the resonator (second term of the equation).

C_0 and the motional values can be now defined:

$$C_0 \geq \zeta_{piezo} \cdot \frac{Lb}{t_{piezo}}$$

For the next computation of the report, we consider the capacitor of the background noise as:

$$C_0 \approx 100 \cdot \zeta_{piezo} \cdot \frac{Lb}{t_{piezo}}$$

t_{piezo} is the thickness of the piezoelectric and ζ_{piezo} is the dielectric constant of the piezoelectric. $C_m = 7.35 \cdot \frac{d_{13}^2 z_{offset}^2 LYb}{t_{total}^3}$

$$L_m = 0.136 \cdot \frac{t_{total}^3}{\omega_0^2 d_{13}^2 z_{offset}^2 LYb} = \frac{1}{\omega_0^2 \cdot C_m}$$

$$R_m = 0.136 \cdot \frac{t_{total}^3}{\omega_0 Q d_{13}^2 z_{offset}^2 LYb} = \frac{1}{\omega_0 \cdot Q \cdot C_m}$$

(Villanueva, Equivalent Circuit of a Flexural Beam Resonator, 2015)(Enz & Kaiser, 2002)

2.3 Reflective wave

For this project, the concept of reflective wave and wave propagating in the circuit has to be introduced. In fact, the current and the voltage in an electric circuit can be considered as a propagating electromagnetic wave. The electric field can be compared to the voltage and the magnetic field can be compared to the current. When a 50 Ω transmission line or any device which has 50 Ω impedance is used, it doesn't necessarily mean that the impedance of this line/device is 50 Ω. It means that the ratio between the propagating electric field and the propagating magnetic field inside the line is 50.

What is happening when a 50 Ω line is connected with a 100 Ω component? First, we see that the electromagnetic wave has to respect the 50 Ω ratio:

$$\frac{\|\vec{E}\|}{\|\vec{H}\|} = \frac{50}{1} = 50$$

Then, when the electromagnetic wave reaches the 100 Ω component, the wave has to respect the 100 Ω ratio:

$$\frac{\|\vec{E}\|}{\|\vec{H}\|} = \frac{100}{1} = 100$$

We see that there is a contradiction because the electric field gained a value of twice of its initial value. In term of energy, it is not possible. To solve this contradiction, a reflective wave has to be considered. Consequently, to respect the 100 Ω ratio, the electric field and the magnetic field become:

$$\frac{\|\vec{E}\|}{\|\vec{H}\|} = \frac{50}{0.5} = 100$$

The magnetic field has decreased in order to respect the 100 Ω ratio. It means that the rest has been reflected. However, the phenomenon is more complex because after the reflection, the wave has to respect again the 50 Ω impedance ratio. So some electric field is also reflected.

The reflection coefficient is:

$$\Gamma = \frac{Z_{load} - Z_{source}}{Z_{load} + Z_{source}}$$

If the source has a 50 Ω impedance, to avoid the reflective wave, the load has to have also an impedance of 50 Ω. If it is not the case, a matching of the two impedances is needed.

The voltage standing wave ratio (VSWR) is a parameter which defines how well the two impedances are matched.

$$VSWR = \frac{1 + \|\Gamma\|}{1 - \|\Gamma\|}$$

If the VSWR parameter is equal to one the impedances are perfectly matched.

The percentage of power delivered from the source to the load impedance is:

$$P_{transmitted} = 100 \cdot (1 - \|\Gamma\|^2)$$

Again, it can be seen that the reflection coefficient has to be equal to zero to maximize the delivered power.

(Locher, 2009)

2.4 Impedance Matching

In some cases, the source and the load are not matched. As we have seen in the previous chapter, the impedance of the load and the impedance of the source have to be as close to equal as possible to have a good matching. To match the source and the load, some components between the source and the load can be implemented to change the impedances. Figure 4 shows a potential schematic of the impedance matching. There are many ways to do the impedance matching. In this project, we are interested in using two types of matching: LC-Matching and Transformer-Matching.

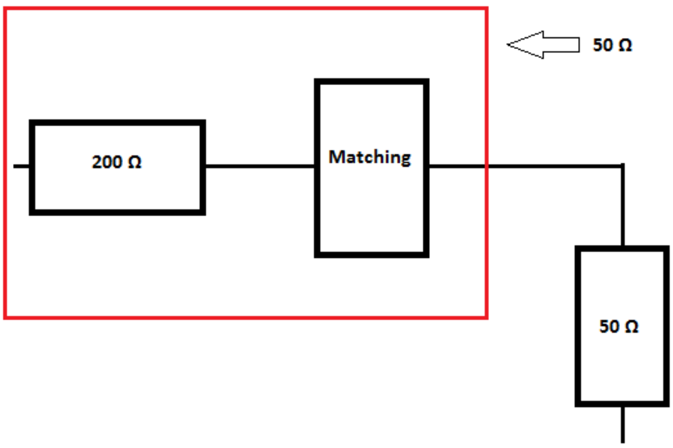

Figure 4 - Impedance matching

LC-Matching

The LC-Matching adds an LC circuit between the load and the source. One capacitor and one inductor are used. One of the components is put in parallel (linked to the ground) and the other is put in series between the impedances to match. The values of the inductor and the capacitor are computed in such a way so that the two impedances are matched. A website computes the value of the LC network to have the best match as possible:(Wetherell, 1997)

The website gives different possible designs that we can put between the source and the load with their specific LC values. The LC values are calculated in such a way that the resistance of the load (generally 50 Ω) has the same resistance as the impedance of the source combined with the impedance of the LC-Matching.

The impedance of the source and the load have to be set beforehand. It is also important to set the working frequency. In this calculation, the Q factor was not considered and was set to a random value (There is no effect on the chosen design). The most interesting and easy designs to implement are the highpass and the lowpass. No significant difference has been noticed between the highpass matching and the lowpass matching.

Transformer Matching

The transformer changes the ratio between the voltage and the current. By analogy, it is changing the ratio between the electric field and the magnetic field of the propagating wave. Thus, it is a very suitable way to match two impedances.

Figure 5 shows how the matching is done between the impedances Z_L and Z_S.

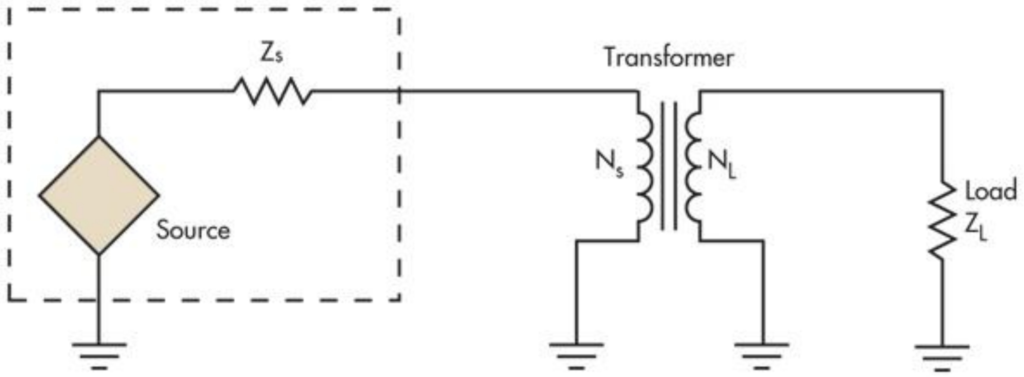

Figure 5 - Transformer Matching

The basic equation of an ideal transformer are:

$$T = \frac{N_S}{N_L} = \frac{U_S}{U_L} = \frac{I_L}{I_S}$$

(Perriard & Köchli, 2014)

To match the impedances Z_L and Z_S:

$$\frac{\|Z_S\|}{\|Z_L\|} = \left(\frac{N_S}{N_L}\right)^2 = T^2$$

T is the transformer ratio (look out: according to some softwares or datasheets, the value of the transformer ratio can be $\frac{1}{T}$), N_i is the number of windings, I_i is the current at the corresponding impedance Z_i , U_S is the voltage between the impedance Z_S and the transformer and U_L is the voltage on the impedance Z_L.(Frenzel, 2011)

3. Evolution of the Project and Results

3.1 Values of the RLC resonator circuit and background noise capacitance

As seen in the theory, the main part of the chip is a resonator (R, L and C circuit) with a background noise capacitance C_0 in parallel of the resonator. Table 1 shows the value of the flexural beam.

Name of the parameter	**Parameter**	**Value and range**
Piezoelectric coefficient	d_{13}	$2.5\ [pmV^{-1}]$
Young modulus	Y	$300\ [GPa]$
Length of the beam	L	$10\ [\mu m]$
Width of the beam	b	$0.5 - 5\ [\mu m]$
Total thickness of the beam	t_{total}	$115\ [nm]$
Offset distance of the beam	z_{offset}	$32.5\ [nm]$
Resonance frequency	f_0	$0.8\ [Hz]$
Thickness of the piezoelectric	t_{piezo}	$50\ [nm]$
Q-factor	Q	1000
Dielectric constant	ζ_0	$8.854187\ [pFm^{-1}]$
Relative dielectric constant of the piezoelectric	$\zeta_{relative,piezo}$	9
The C_0 of the background noise is 10 to 100 times the geometrical capacitance		

Table 1 - Dimension of the flexural beam

The different values of the capacitance of the background noise are shown in Table 2.

C_0	$10 \cdot C_0$	$100 \cdot C_0$
$80\ [nF]$	$800\ [nF]$	$8\ [pF]$

Table 2 - Capacitance of the background noise

Motional values of the RLC circuit of the resonator calculated on Matlab as a function of two different widths of the beam are shown in Table 3.

	R_m	L_m	C_m
$b = 0.5\ [\mu m]$	$4157\ [k\Omega]$	$827.0\ [H]$	$0.0479\ [fF]$
$b = 5[\mu m]$	$415.7\ [k\Omega]$	$82.70\ [H]$	$0.479\ [fF]$

Table 3 – Motional values of the RLC circuit of the resonator

For every simulation, the background noise capacitance is 100 times the geometrical capacitor and a width of 5 µm will be considered.

3.2 Quite Universal Circuit Simulator

Every circuit simulation for this project is done on the Quite Universal Circuit Simulator (QUCS). For the following simulation, the S-parameter is plotted and is normalized at 1. To analyze the spectrum of the S-parameter, two alimentation sources with an impedance of 50 Ω are put on both sides (source and detector) of the circuit. The S_{ii} expresses the reflective coefficient and the S_{ij} expresses the transmission coefficient of the propagating wave between the “i” and the “j” alimentation sources. We will be only interested in the transmission coefficient and maximizing it. There is no need to set an alimentation voltage because the S-parameter doesn't depend on the value of the voltage. The design and the schematic of the electric circuit with the required component for the simulation are shown in the next chapter.

(Jahn & Borrás, 2007)

3.3 Resonator LC-Matching

For all the LC-Matching, these following steps are done: First, the impedance Z_i of the resonator is defined. Z_i has to be matched with a 50 Ω impedance which is the impedance of the detector. Then, to find the LC-matching, the website that computes the value of L and C for the impedance matching is used. The Z_i is considered as the source impedance and the 50 Ω resistor is considered as the load impedance. The lowpass matching network design is taken.

First, only the resonator will be matched. The resonator's impedance is:

$$Z_m = R_m + j \cdot \left(\omega_0 L_m - \frac{1}{\omega_0 C_m}\right)$$

Knowing that: $\omega_0 = 2\pi f_0$

$$Z_m = 41.57\ [k\Omega] = R_m$$

We note that there is no imaginary part of the impedance because the resonator is working at the resonance frequency.

The value of the inductor and the capacitor are shown in Table 4. These first previous steps are going to be explained only for this simple case.

Inductor	*Capacitor*
$0.912\ [mH]$	$43.4\ [pF]$

Table 4 - LC-matching of the Z_m

Figure 6 shows the circuit with the correct value on QUCS. The two alimentation sources are placed at both sides of the RLC resonator to do the simulation of the S-parameter.

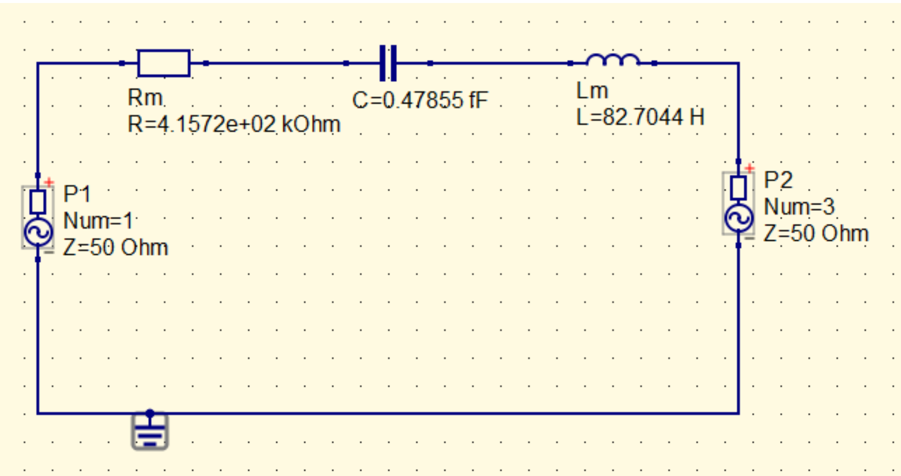

Figure 6 - Z_m circuit

Figure 7 shows the resonator with the computed value of the matching by the website.

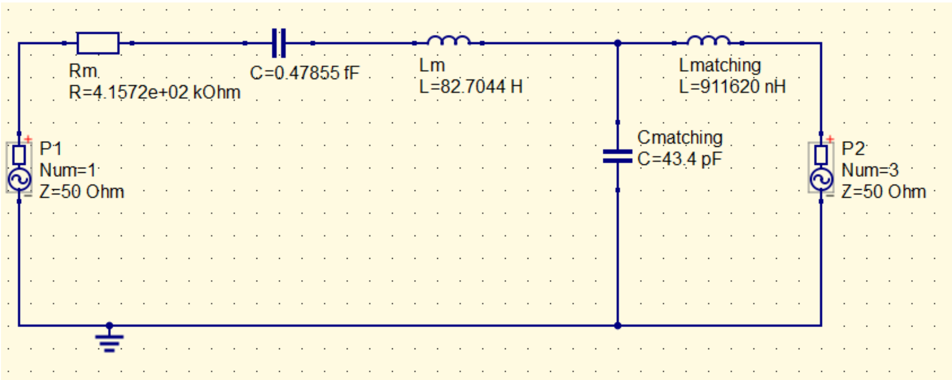

Figure 7 - Zm circuit with the LC-matching

Figure 8 shows the plot of the transmission S-parameter of the resonator as a function of the frequency. There is a peak of about $2.5 \cdot 10^{-4}$ at a resonance frequency of $0.8\ MHz$. Figure 9 shows the plot of the transmission S-parameter of the resonator with the matching as a function of the frequency. The peak has a value more than 10^{-2}.

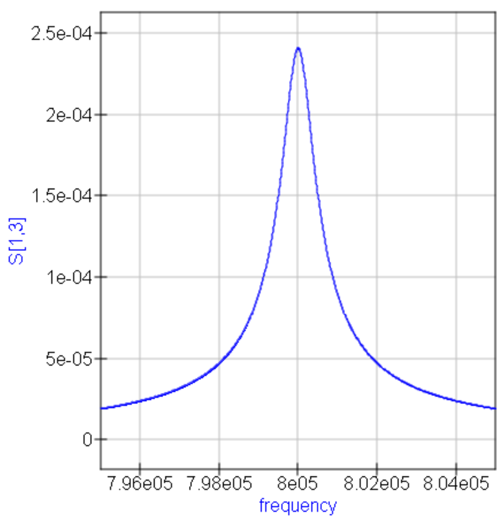

Figure 8 - S-parameter of the resonator

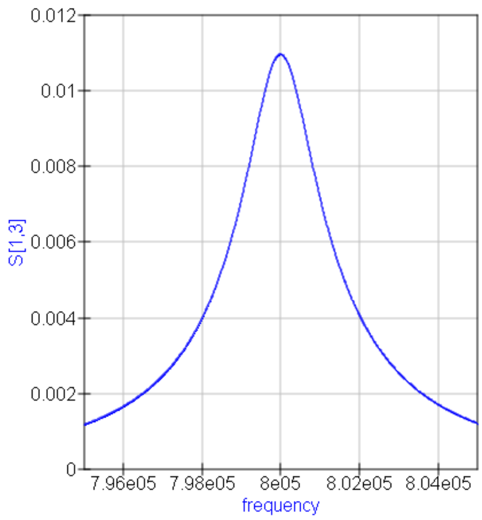

Figure 9 - S-parameter of the matched resonator

The signal is almost improved by two orders. The matching is already well realized.

3.4 Resonator with background noise LC-matching

Let's see the S-parameter of the resonator when the background noise is considered. The impedance of resonator with the background noise capacitance is:

$$Z_{r,C_0} = (Z_m^{-1} + j\omega_0 C_0)^{-1} = 1.49 - j24.9\ [k\Omega]$$

The value of the LC-matching of Z_{r,C_0} is calculated on the website and is shown in Table 5.

Inductor	*Capacitor*
$0.909\ [mH]$	$35.6\ [pF]$

Table 5 - LC-matching of $Z_{r,C0}$

The circuit of $Z_{r,C0}$ with the matching is shown in Figure 10.

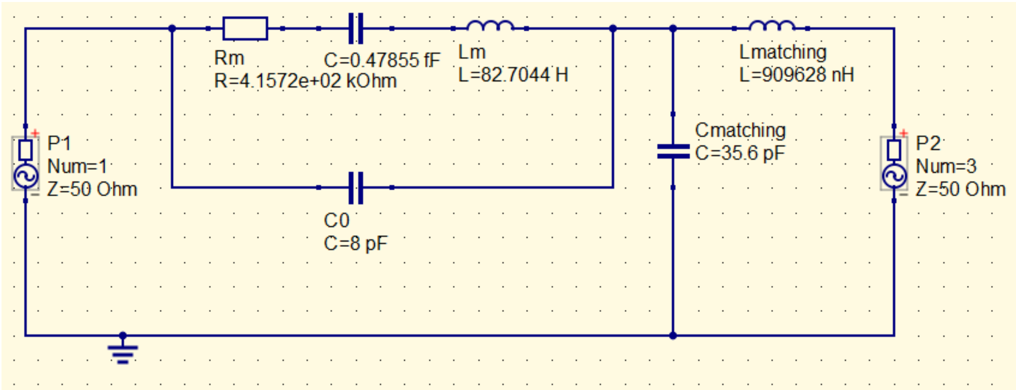

Figure 10 - $Z_{r,C0}$ circuit with the LC-matching

The S-parameter of the resonator with C_0 is shown in Figure 11 and Figure 12.

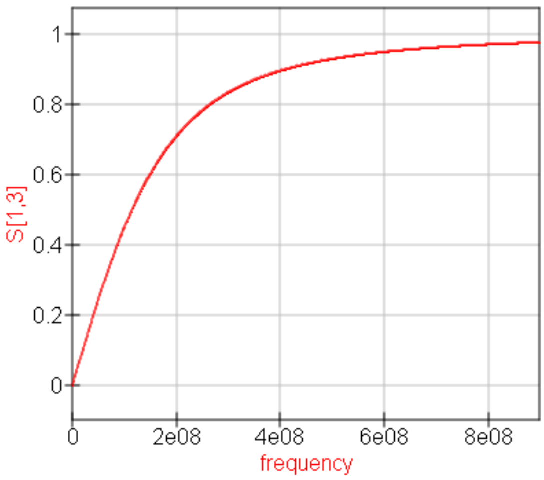

Figure 11 - S-parameter of the resonator with C0: wide spectrum

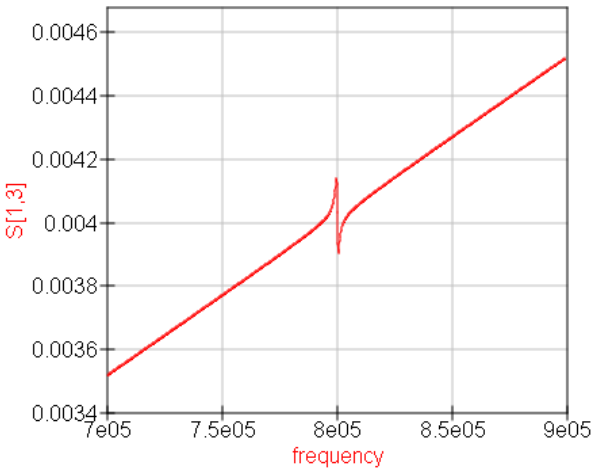

Figure 12 - S-parameter of the resonator with C0: Focused on the resonance frequency

It can be seen that the capacitance of the background noise acts like a highpass filter (Figure 11) and the peak is situated on the slope of this highpass filter at the resonance frequency of the resonator (Figure 12).

Figure 13 shows the transmitted S-parameter of the resonator with C_0 and with the matching.

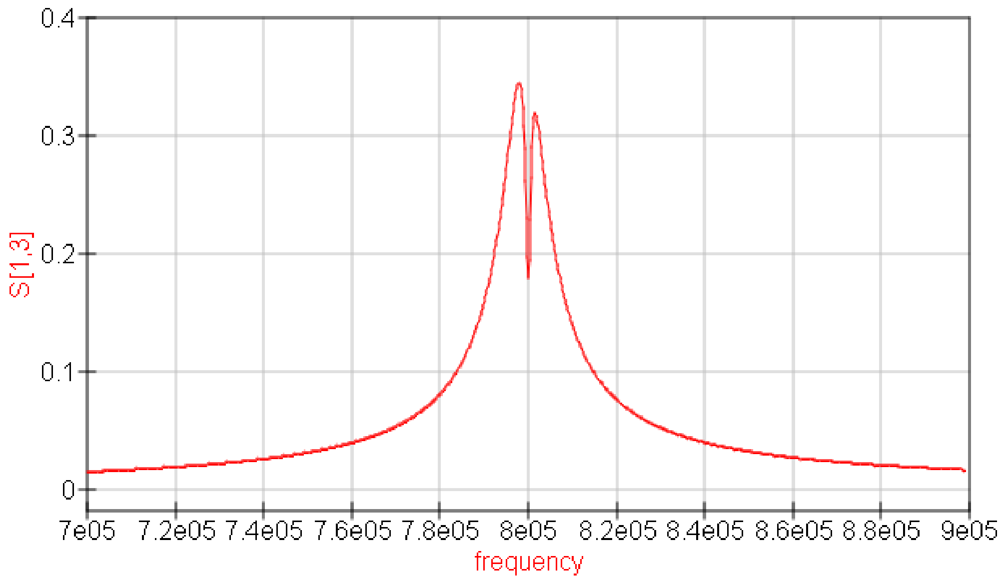

Figure 13 - S-parameter of the matched resonator with C_0

The plot of the S-parameter shows that the LC-matching does amplify the resonance frequency. Therefore the peak shown in Figure 12 is locally amplified and the result is shown in Figure 13. This is why we see a wide peak combined with a reversed small one. Of course, this is not suitable to measure anything.

The matching of this resonator with the LC-matching values of the simple resonator (Table 4) has been tried with no conclusive results.

3.5 Cancellation of the background noise with LC-matching

Clearly, the problem in the previous chapter is the background noise capacitance. A way to cancel it is to add another line parallel to the chip which does the reversed effect of the background noise capacitance. There are different ways to do it:

Phase shifter with a Capacitor

The additional line uses a phase shifter that shifts the signal by an angle of 180 degrees. For the simulation, a phase shifter is used but in practice, a power splitter is going to be used. The power splitter creates another signal with a shifted angle of 180 degrees. A capacitor of the same value as the background noise capacitance is added in series. Thus, a background noise signal with the same amplitude but with an opposite angle is created and will cancel the background noise.

Figure 14 shows the circuit with the phase shifter and the capacitor in parallel of the resonator with a following LC-matching.

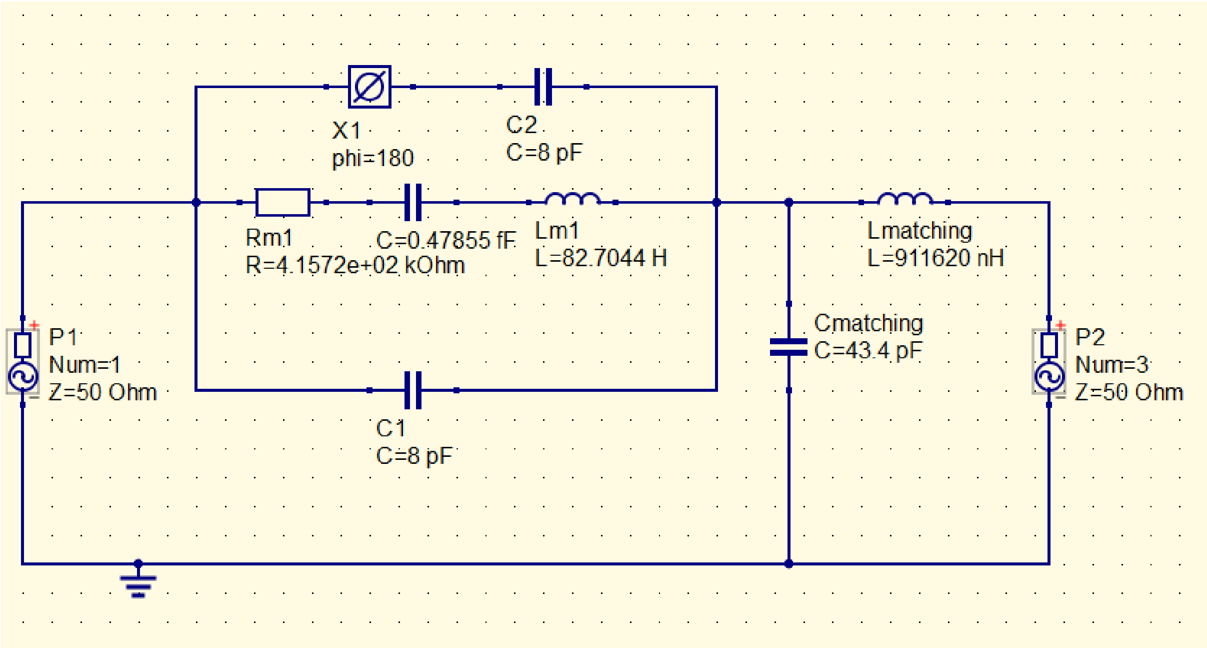

Figure 14 – Z_{tot} circuit with a LC-matching

The total impedance of the chip with the background cancellation is:

$$Z_{tot} = \left(Z_{r,C_0}^{-1} + j\omega_0 C_{line}\right)^{-1} = 0.373 - j12.4\ [k\Omega]$$

If we consider that the background noise is cancelled, the LC-matching of only Z_m is tried (Values of the matching are shown in Table 4).

Figure 15 and Figure 16 show the comparison of the S-parameters between the unmatched signal and the LC-matched signal with Z_m.

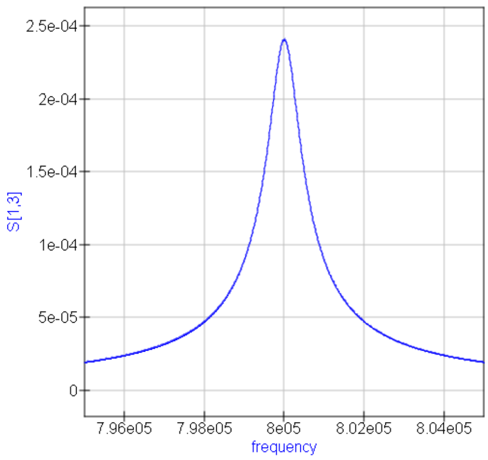

Figure 15 - S-parameter of the chip with background cancellation

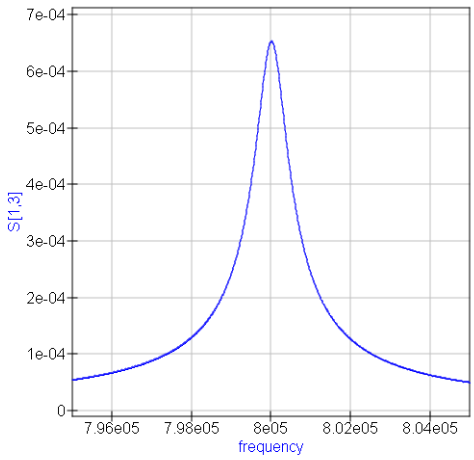

Figure 16 - S-parameter of the chip with background cancellation LC-matched with Zm

First, we see that the background noise is cancelled. Additionally, there is an improvement of more than twice than the unmatched signal but it is not sufficient. So the LC-matching of Z_{tot} is tried.

The value of the LC-matching is shown in Table 6

Inductor	*Capacitor*
$874[mH]$	$28.7\ [pF]$

Table 6 - LC-matching's value of Z_{tot}

Figure 17 shows the plot of the S-parameter in function of the frequency with the LC-matching of Z_{tot}.

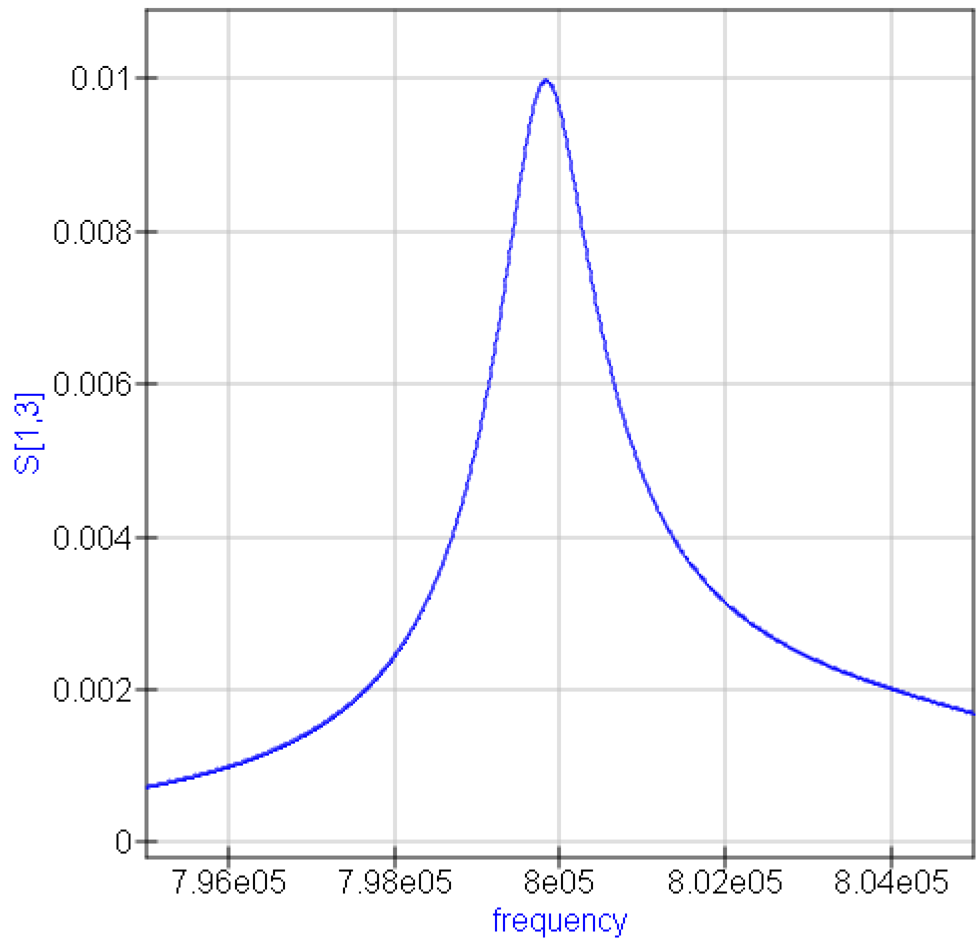

Figure 17 - S-parameter of the chip with background cancellation LC-matched with Ztot

The improvement is about 40 times better. It is clearly better than the matching with Z_m.

Inductor in parallel

With this method, the goal of the line is to cancel the impedance generated by the background capacitance with an inductor in parallel. The overall impedance is:

$$Z_L = \left(Z_m^{-1} + j \left(\omega_0 C_0 - \frac{1}{\omega_0 L_{line}} \right) \right)^{-1}$$

So we want that the contribution of C_0 is cancelled:

$$L_{line} = \frac{1}{\omega_0^2 C_0} = 5\ [mH]$$

The overall impedance becomes:

$$Z_L = Z_m$$

Figure 18 shows the circuit with the inductor in parallel and with the LC-matching.

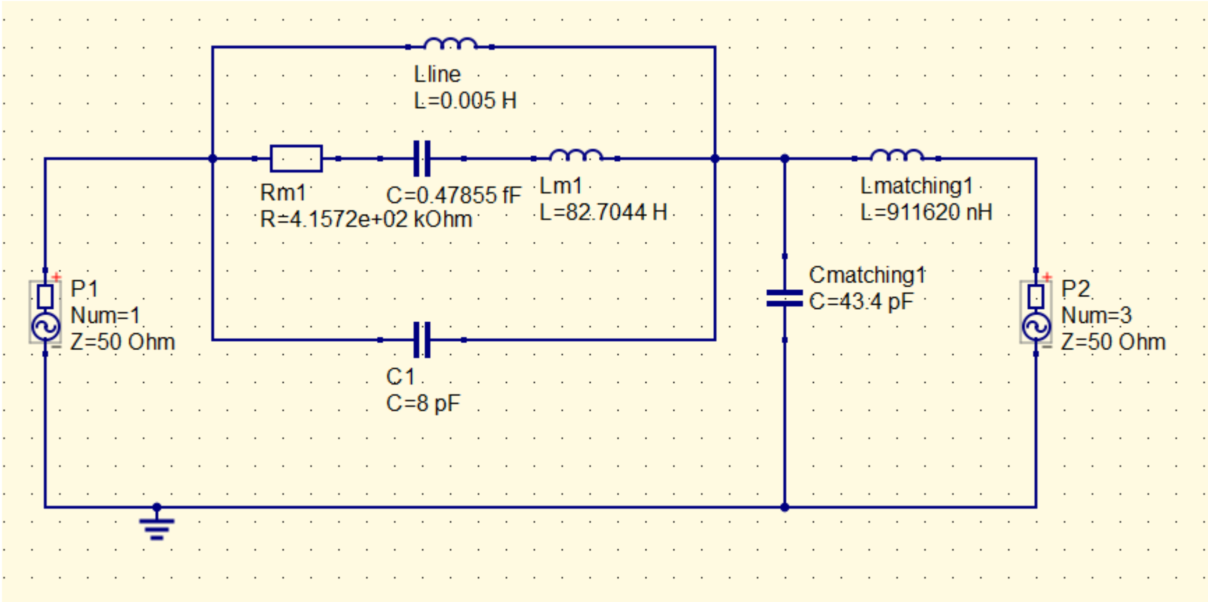

Figure 18 - Circuit with the inductor cancellation line and a LC-matching

To match this circuit, the impedance $Z_L = Z_m$ is used. The values of the LC-matching are shown in Table 4.

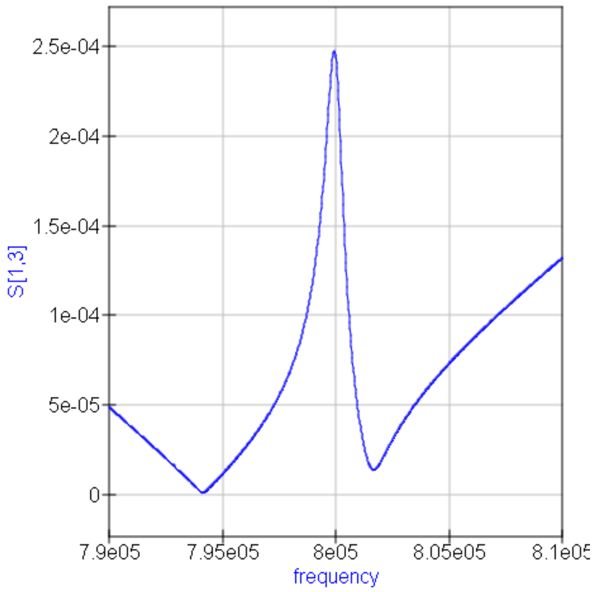

Figure 19 - Unmatched Z_L

Figure 20 - LC-Matched Z_L

The improvement of the signal is about 40 times better.

We can see that the cancellation of the background noise works well and if the right LC-matching is chosen, it can improve the signal of about 40 times.

(Villanueva, Background Noise Cancellation line of a flexural beam resonator, 2015)

3.6 Cancellation of the background noise with the transformer matching

Problem of the LC-matching

The problem of the LC-Matching is its small bandwidth. We see that the LC-matching works very well with QUCS because all the parameters are exactly defined. Therefore the LC-Matching is convenient for some well known circuits. However, in practice and in the project, some parameters of the circuit are approximated (the resonance frequency and background noise capacitance of the flexural beam). Additionally, the phase shifter has an unknown impedance. Consequently the LC-matching is not suitable for this project. The bandwidth of the transformer is wider. Therefore, the transformer matching is more convenient for the project and will be used for the next simulations. Note that a perfect match is never able to be realized since the overall impedance of the resonator is not exactly defined.

To demonstrate the limitations of the LC-Matching, the resonance frequency has been switched to 8 MHz. The motional RLC elements change and are shown in Table 7. The background capacitance remains the same.

	R_m	L_m	C_m
$\omega_0 = 8[MHz]$	$41{,}57\ [k\Omega]$	$0.8270\ [H]$	$0.479\ [fF]$

Table 7 - Motional RLC value of the resonator at 8 MHz

The matching is tried on this resonator with the background cancellation with the capacitor and the phase shifter design (Figure 14). But instead of using the right matching, the matching used previously for the resonator at 0.8 MHz is applied. So the values of Table 6 have been chosen for the simulation.

Figure 21 and Figure 22 show the plot of the S-parameter at 8 MHz. The unmatched is on the left and the “LC-matched” at 0.8 MHz is on the right.

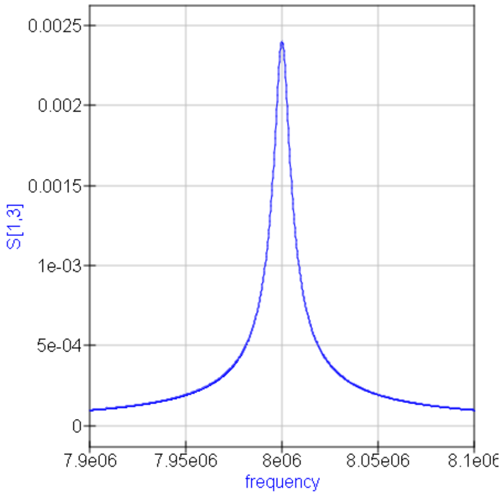

Figure 21 - S-parameter of Ztot at 8 Mhz

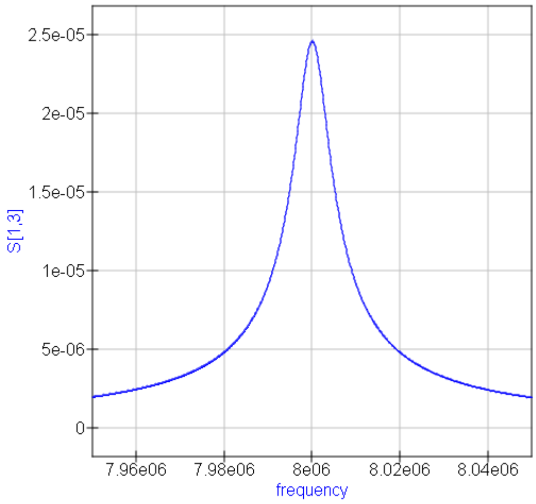

Figure 22 - S-parameter of Ztot at 8 MHz with LC-matching at 0.8 MHz

It can be seen that the circuit is not matched at all and is now worse than being unmatched.

Transformer matching of Z_{tot}

Instead of an inductor and a capacitor, a transformer is used. The transformer is placed between the chip and the alimentation source. The circuit is shown in Figure 23.

The transformer ratio is:

$$T = \sqrt{\frac{\|Z_{tot}\|}{\|Z_L\|}} = \sqrt{\frac{12'453}{50}} = 15.78$$

The value of the transformation ratio is set at 16 for the simulation.

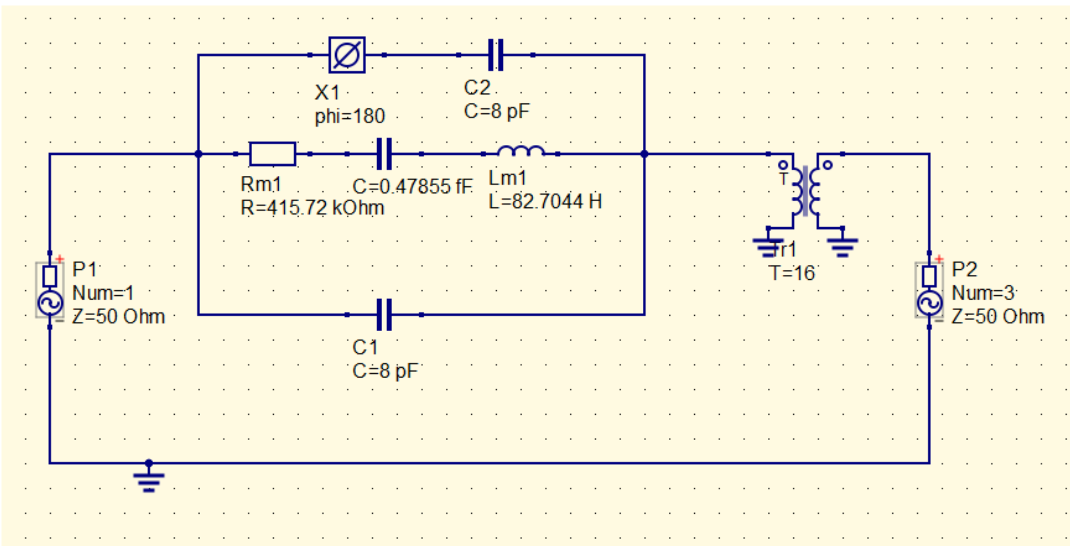

Figure 23 - Circuit of Z_{tot} with the transformer matching

Figure 24 and Figure 25 compare the S-parameter value of the chip with the S-parameter value of the chip with the matching.

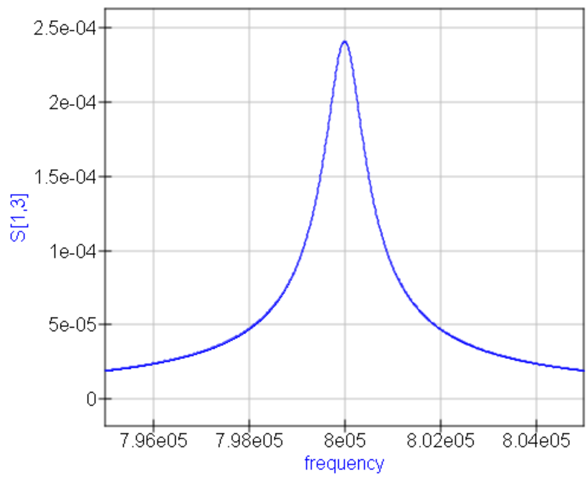

Figure 24 - S-parameter of Ztot

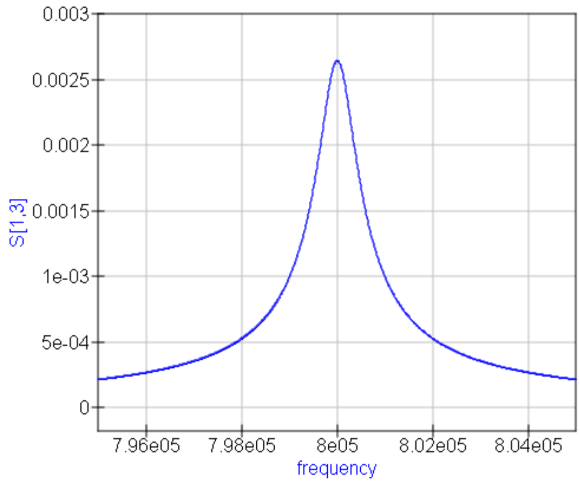

Figure 25 - S-parameter of Ztot with T-matching

There is an improvement of a factor 10 when the chip is matched with the transformer.

Bandwidth of the transformer

We have showed with an example previously that the LC-matching has a small bandwidth. Let's test now the bandwidth of the transformer.

As previously, the equivalent circuit is set with a resonance frequency at 8 MHz (see Table 7). The background noise and the cancellation line with the capacitor and the phase shifter are considered (see Figure 23).

Figure 26 and Figure 27 show the plot of the S-parameter at 8 MHz. The unmatched is on the left and the "Transformer-matched" at 0.8 MHz is on the right.

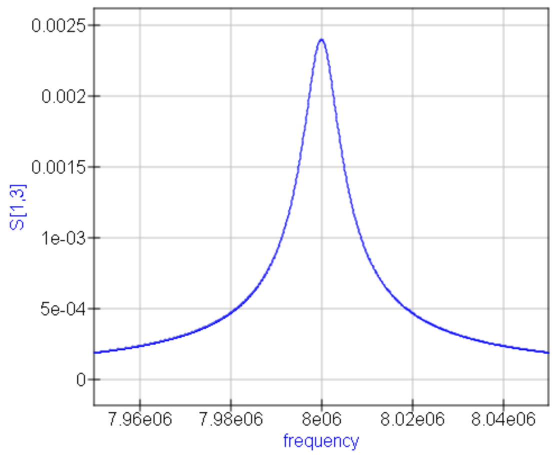

Figure 26 - S-parameter of Ztot at 8 MHz

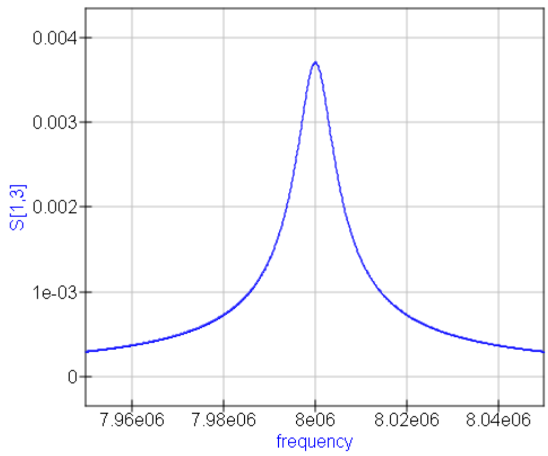

Figure 27 - S-parameter of Ztot at 8 MHz with T-matching at 0.8 MHz

We can see that there is still an improvement of a factor two even if the frequency has been switched. This is why it is more convenient to use a transformer.

3 -Transformers matching of Z_{tot}

A matching can be also done on the left side of the circuit. Figure 28 shows the circuit of the matching of the chip.

The transformation ratio on the left side has to match the impedance from the 50 Ω impedance of the power splitter and the impedance of the chip:

$$T = \sqrt{\frac{\|Z_S\|}{\|Z_{tot}\|}} = \sqrt{\frac{50}{12'453}} = 0.0634$$

The value of the transformation ratio is set at $\frac{1}{16} = 0.0625$ for the simulation.

Figure 28 shows the circuit of the 3 Transformers-matching.

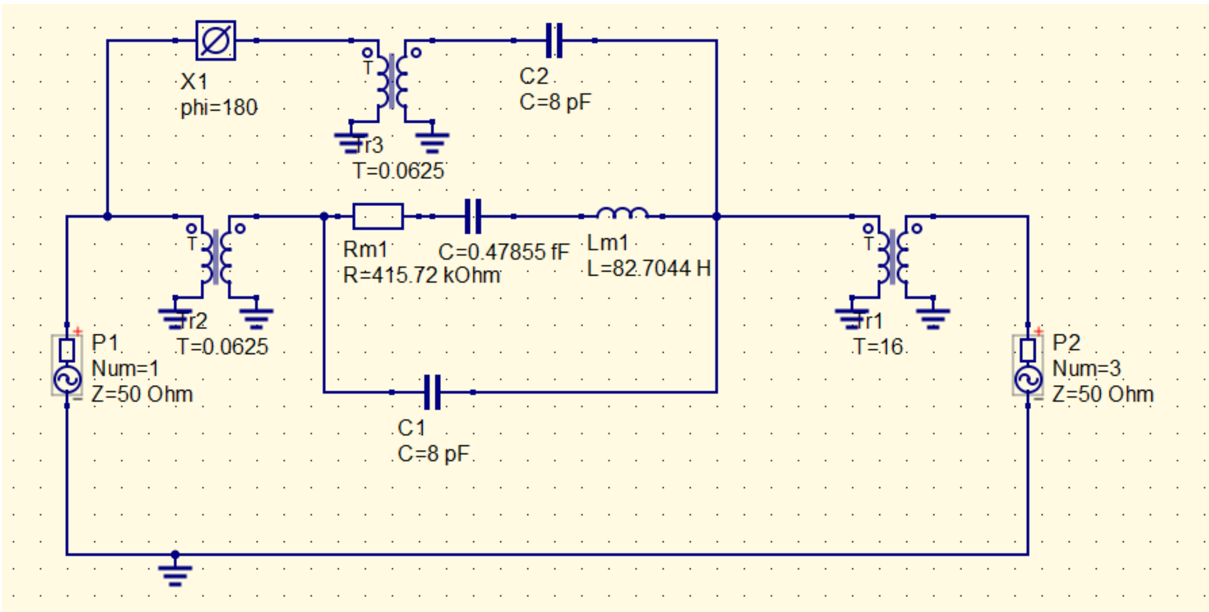

Figure 28 - Circuit of Z_{tot} with 3 transformers matching

The matching is unnecessary between the source and the power splitter. Since the power splitter also has an impedance of 50 Ω. To do the matching between 50 Ω impedance of the power splitter and Z_{tot}, two additional transformers are used. One for the chip and one for the cancellation line (see Figure 28).

Figure 29 shows the S-parameter of the chip in function of the frequency and Figure 30 shows the S-parameter of the chip with the 3T-matching in function of the frequency.

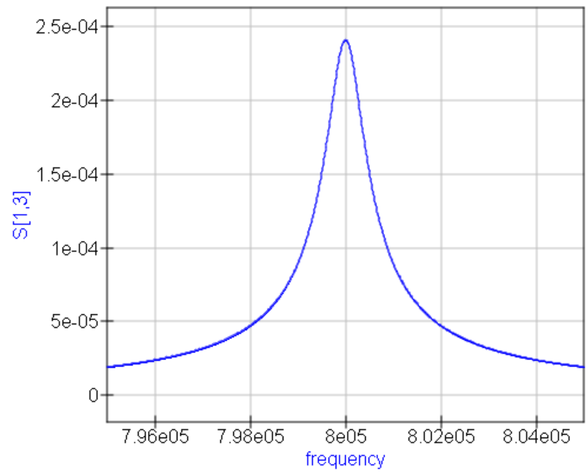

Figure 29 - S-parameter of Ztot

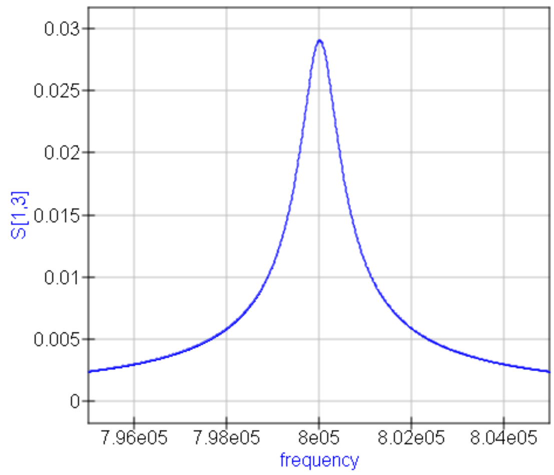

Figure 30 - S-parameter of Ztot with the 3T-matching

The improvement is more than 100 times better. It is better than the single transformer matching.

Another way to cancel the background noise capacitance is to put in parallel of the chip a phase shifter with an attenuator. We could have a nice plot of the S-parameter with a good matching. Unfortunately, the file has been lost and the data couldn't be found again. Some other simulations of this design have been tried but weren't successful.

4. Choice of design and Components

We have to consider that the impedances on QUCS are ideal and the components on the market are real. To define which design of chapter 3 is the best, some criteria have been established:

- The nominal value should correspond to the value of the value of the simulation.
- The frequency range should be from $0.5\ MHz$ to $20\ MHz$
- Since the values of all the motional elements of the resonator are not exactly defined and we don't know the value of the background capacitance (it can be 10 or 100 times the geometrical capacitance between the two electrodes), some variable components need to be chosen.
- The variable components have to be tunable with a screwdriver due to an easier handling of the device.
- The components need to be "Surface Mounted Devices" (SMD) because a PCB is used.
- The bandwidth of the matching cannot be small.

4.1 LC-matching or Transformer matching

As said in the chapter "Problem of the LC-matching", the LC-matching is not going to be chosen due to its small bandwidth. So the discuss will be only about the transformer and the choice of a right cancellation line. The matching with three transformers is chosen because it has a higher transmission coefficient at the resonance frequency than the single transformer matching.

4.2 Inductor cancellation line

The cancellation line with the inductor is not doable. As seen in the chapter “Inductor in parallel” the value of the inductor is $L_{line} = 5\ [mH]$. On the market, higher the nominal value of a variable inductor is, lower its frequency test is. Unfortunately, for these kind of variable inductors, the frequency test is not higher than $0.79\ MHz$.

4.3 Phase shifter with attenuator cancellation line

No variable attenuators which can be tuned by a screwdriver have been found. All of them are tuned digitally or with a DC-voltage which is not convenient for the project. No variable phase shifters which are SMD have been found.

4.4 Phase shifter with capacitor cancellation line

For this design, a phase shifter of 180 degrees phase shift angle is needed. A power splitter is used. Some SMD power splitter is found on the market. The right value of the capacitor has to be set. A lot of different variable capacitor can be found on the market. Plus, they can be tuned with a screw driver.

The design with three transformers matching and the cancellation line with the phase shifter and the variable capacitor seems to be the most convenient according to the available components on the market (see chapter “3 -Transformers matching of Z_{tot}”).

4.5 List of components

The website and the data sheet of the components are listed at the very end of the report.

5. **PCB on Altium**

According to the datasheets of the entire three components, the schematic library and the PCB design of the components has been done. The entire circuit design has been done on Altium. Figure 31 shows a first try and the design can be surely improved. The design of the transformers should be different.

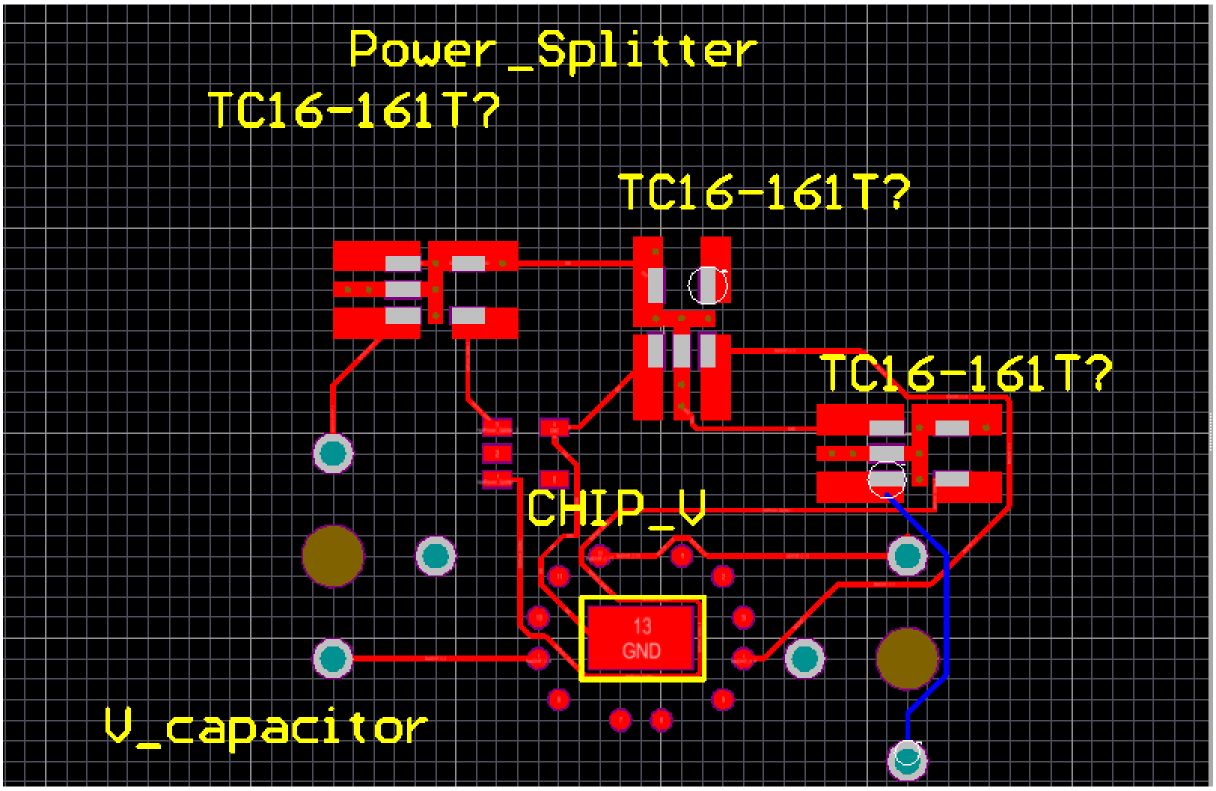

Figure 31 - First design on Altium

6. Conclusion

In the theory, we have been able to define the values of the RLC resonator with the background noise capacitance and how the flexural beam works. The concept of a propagating wave has been explained and allowed to conclude how to have the best matching as possible. The values of the two impedances have to be as close as possible (most of the time 50 Ω). We have been able to define how to match with an inductor and a capacitor and how to match with a transformer.

In the semester project, Quite Universal Circuit Simulation was a very useful software for plotting the S-parameter, thus to analyse different designs and different types of matching.

The LC-matching of the resonator without the background noise has been done. It showed a peak at the wanted resonance frequency and the LC-matching has been successful. We noted an improvement of about 2 orders.

To solve the problem of the background noise, a background noise cancellation line has been added in parallel to the chip. Two designs have been proposed. The first design comprises a phase shifter with a phase shift of 180° and a capacitor of the same value as the background noise capacitance are added in series. The second design comprises an inductor which cancels the impedance generated by the background noise capacitance. The LC-matching and the transformer matching have been successfully simulated (improvement of the transmission S-parameter of at least 1 order). Finally an improvement of 2 orders has been successfully simulated with a design of 3 transformers.

We compared the different designs and we note that the disadvantage of the LC-matching is its too small bandwidth. It requires knowing the exact value of all parameters of the resonator and its resonance frequency which is not the case. The best way therefore to do a matching in the case of the flexural beam resonator is to use some transformers. In fact, the transformers have a possibility to match well the impedances even if the values of the elements of the resonator are not well defined. The 3-transformers matching design has been chosen for this project.

According to what is available on the market, the “cancellation line” with a power splitter and a variable capacitor seems to be the most suitable design for the PCB because of the easy handling of a variable capacitor with a screwdriver.

We have proposed a first PCB design of Altium. It can be improved in term of density and visibility.

What about the future steps?

The PCB design has to be improved so that it can be more dense and understandable. For more information about the fabrication of the PCB, Peter Brühlmeier is the responsible of PCB fabrication at the EPFL. The GERBER files of all the layers have to be created and can be uploaded on the website. The useful information is given in the next page in the chapter “References and useful information”.

I would like to thank my supervisor Guillermor Villanueva for the advice and assistance during the whole semester project.

I would like also to thank Kaitlin Howell.

7. References and useful information

7.1 References

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Frenzel, L. (2011, 10 24). *Back to Basics: Impedance Matching (Part 1)*. Consulté le 01 2016, 06, sur Electronic Design: http://electronicdesign.com/communications/back-basics-impedance-matching-part-1

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Locher, B. (2009). *Radio Frequency Design*. Consulté le 01 06, 2016, sur Fourrier Series Tutorial: http://www.fourier-series.com/rf-concepts/index.html

Perriard, Y., & Köchli, C. (2014). Cours de Bachelor Microtechnique: Conversion électromécanique I et II.

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Villanueva, G. (2015). Equivalent Circuit of a Flexural Beam Resonator.

Wetherell, J. (1997, 10 30). *Impedance Matching Network Designer*. Consulté le 01 06, 2016, sur University of San Diego: http://home.sandiego.edu/~ekim/e194rfs01/jwmatcher/matcher2.html

7.2 Useful information

Information for PCB fabrication :

http://sti-ateliers.epfl.ch/page-19946.html

Upload of the GERBER files for the PCB fabrication

http://aci-commandes.epfl.ch

8. **Datasheets**

Power Splitter:

Datasheet:

Surface Mount, Micro-Miniature

Power Splitter/Combiner

SBTCJ-1W+

2 Way-180° 50Ω 1 to 750 MHz

CASE STYLE: AT790
PRICE: $ 5.95 ea. QTY. (10)

+RoHS Compliant
The +Suffix identifies RoHS Compliance. See our web site for RoHS Compliance methodologies and qualifications

Maximum Ratings

Operating Temperature	-40°C to 85°C
Storage Temperature	-55°C to 100°C
Power Input (as a splitter)	0.5W max.
Internal Dissipation	0.125W max.

Permanent damage may occur if any of these limits are exceeded.

Pin Connections

SUM PORT	6
PORT 1	1
PORT 2	3
GROUND	2,4
NOT USED	5

Features

- low insertion loss, 0.7 dB typ.
- good isolation, 23 dB typ.
- good VSWR, 1.25 typ. all ports
- small size, 0.15X0.15"X0.15"
- temperature stable, LTCC base
- low cost
- protected by US Patent, 6,806,790

Applications

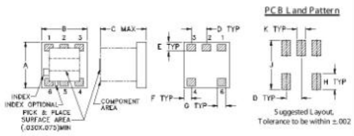
- cellular
- UHF/VHF receivers/transmitters
- protected by US Patent, 6,806,790

Electrical Specifications

FREQ. RANGE (MHz)	ISOLATION* (dB)						INSERTION LOSS (dB) ABOVE 3.0 dB						PHASE UNBALANCE (Degrees)			AMPLITUDE UNBALANCE (dB)		
	L		M		U		L		M		U		L	M	U	L	M	U
f_L-f_U	Typ.	Min	Typ.	Min	Typ.	Min	Typ.	Max.	Typ.	Max.	Typ.	Max.	Max.	Max.	Max.	Max.	Max.	Max.
1-750	23	20	22	20	24	20	0.6	1.7	0.6	1.2	0.9	1.8	3	7	10	0.2	0.4	0.9

L = low range [f_L to 10 f_L] M = mid range [10 f_L to f_U/2] U= upper range [f_U/2 to f_U]
*Isolation, 17 dB min. at 1 - 3 MHz.

Outline Drawing



Outline Dimensions (inch / mm)

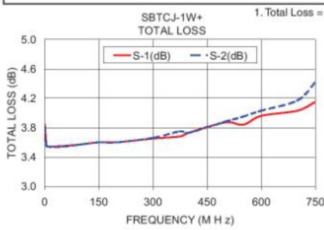
A	B	C	D	E	F
.150	.150	.150	.050	.030	.025
3.81	3.81	3.81	1.27	0.76	0.64

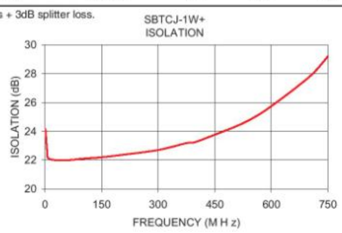
G	H	J	K	wt
.028	.050	.160	.030	grams
0.71	1.27	4.06	0.76	0.10

Typical Performance Data

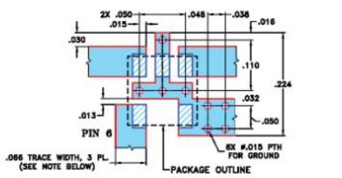
Frequency (MHz)	Total Loss[1] (dB) S-1	Total Loss[1] (dB) S-2	Amplitude Unbalance (dB)	Isolation (dB)	Phase Unbalance (deg.)	VSWR S	VSWR 1	VSWR 2
1.00	3.84	3.80	0.04	24.15	179.96	1.24	1.15	1.14
5.00	3.57	3.56	0.01	22.73	179.96	1.11	1.07	1.06
10.00	3.54	3.54	0.00	22.11	179.95	1.09	1.05	1.05
50.00	3.55	3.54	0.00	21.98	179.53	1.08	1.04	1.05
100.00	3.57	3.57	0.00	22.10	179.01	1.09	1.04	1.06
150.00	3.60	3.60	0.01	22.20	178.54	1.11	1.05	1.09
200.00	3.60	3.60	0.00	22.35	178.04	1.13	1.06	1.12
300.00	3.65	3.66	0.01	22.70	177.08	1.18	1.09	1.19
375.00	3.68	3.75	0.07	23.18	176.60	1.22	1.13	1.25
400.00	3.73	3.73	0.01	23.26	176.18	1.23	1.14	1.28
500.00	3.87	3.88	0.01	24.27	175.75	1.29	1.21	1.38
550.00	3.84	3.95	0.11	24.90	175.68	1.32	1.25	1.44
600.00	3.96	4.03	0.07	25.74	175.59	1.35	1.30	1.50
700.00	4.03	4.17	0.14	27.76	175.20	1.41	1.41	1.64
750.00	4.15	4.42	0.27	29.19	175.46	1.44	1.46	1.72

1. Total Loss = Insertion Loss + 3dB splitter loss.

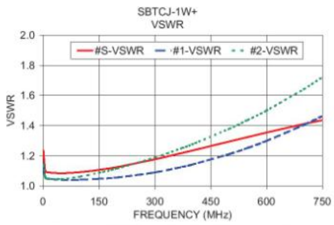




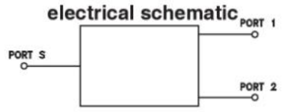
Demo Board MCL P/N: TB-227
Suggested PCB Layout (PL-117)



NOTES: 1. TRACE WIDTH IS SHOWN FOR ROGERS RO4350B WITH DIELECTRIC THICKNESS .030" ± .002"; COPPER: 1/2 OZ. EACH SIDE. FOR OTHER MATERIALS TRACE WIDTH MAY NEED TO BE MODIFIED.
2. BOTTOM SIDE OF THE PCB IS CONTINUOUS GROUND PLANE.
DENOTES PCB COPPER LAYOUT WITH SMOBC (SOLDER MASK OVER BARE COPPER)
DENOTES COPPER LAND PATTERN FREE OF SOLDER MASK



electrical schematic



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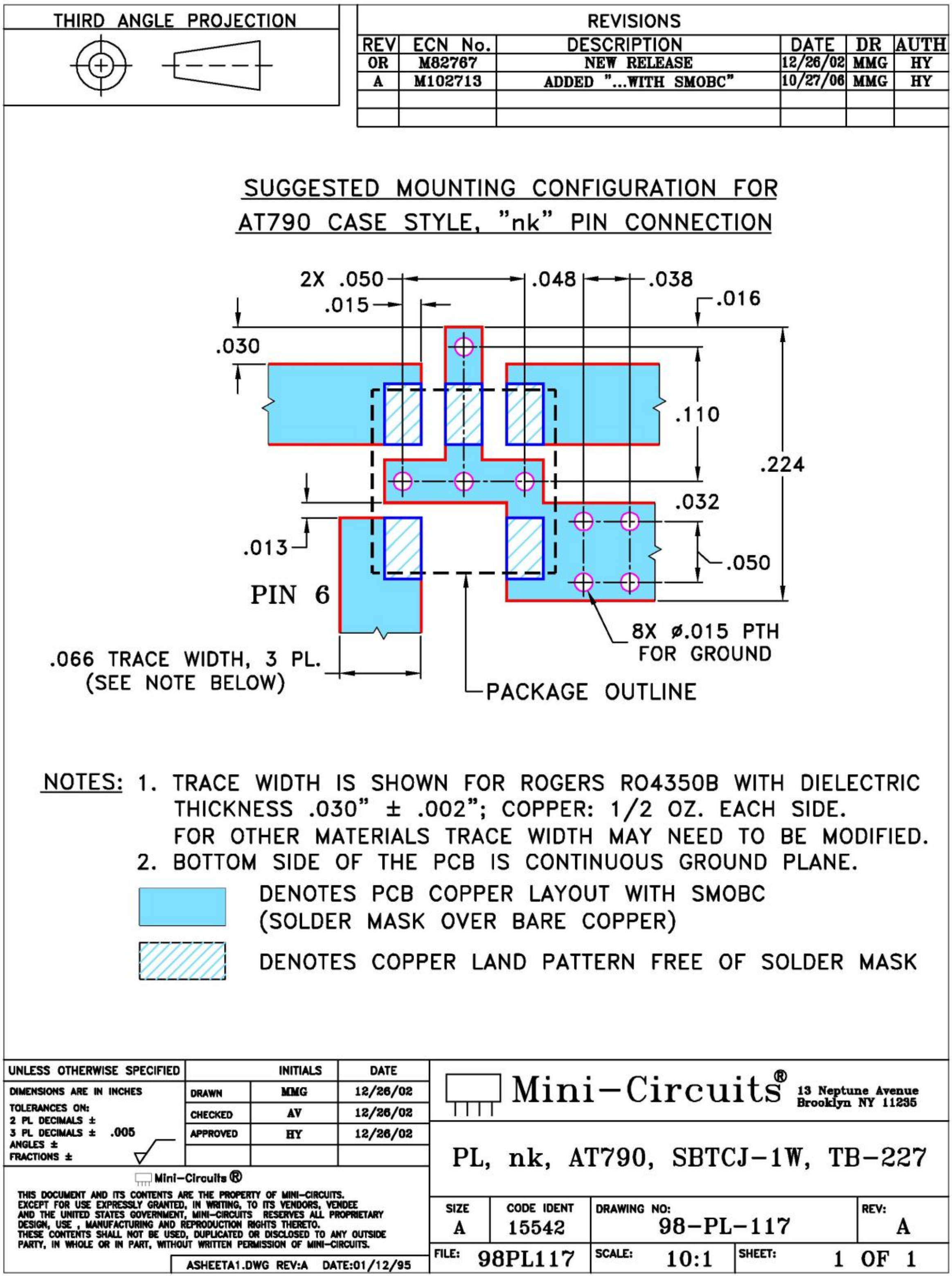
REV. H
M127604
ED-8081A/1
SBTCJ-1W+
DJ/TD/CP/AM
130430

PCB Layout:


THIRD ANGLE PROJECTION
REVISIONS
REV | ECN No. | DESCRIPTION | DATE | DR | AUTH
OR | M82767 | NEW RELEASE | 12/26/02 | MMG | HY
A | M102713 | ADDED "...WITH SMOBC" | 10/27/06 | MMG | HY
SUGGESTED MOUNTING CONFIGURATION FOR
AT790 CASE STYLE, "nk" PIN CONNECTION
2X .050
.015
.048
.038
.016
.030
.110
.224
.032
.013
.050
PIN 6
8X ø.015 PTH
FOR GROUND
.066 TRACE WIDTH, 3 PL.
(SEE NOTE BELOW)
PACKAGE OUTLINE
NOTES: 1. TRACE WIDTH IS SHOWN FOR ROGERS RO4350B WITH DIELECTRIC THICKNESS .030" ± .002"; COPPER: 1/2 OZ. EACH SIDE. FOR OTHER MATERIALS TRACE WIDTH MAY NEED TO BE MODIFIED.
2. BOTTOM SIDE OF THE PCB IS CONTINUOUS GROUND PLANE.
DENOTES PCB COPPER LAYOUT WITH SMOBC (SOLDER MASK OVER BARE COPPER)
DENOTES COPPER LAND PATTERN FREE OF SOLDER MASK
UNLESS OTHERWISE SPECIFIED
DIMENSIONS ARE IN INCHES
TOLERANCES ON:
2 PL DECIMALS ±
3 PL DECIMALS ± .005
ANGLES ±
FRACTIONS ±
INITIALS | DATE
DRAWN | MMG | 12/26/02
CHECKED | AV | 12/26/02
APPROVED | HY | 12/26/02
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PL, nk, AT790, SBTCJ-1W, TB-227
SIZE A
CODE IDENT 15542
DRAWING NO: 98-PL-117
REV: A
FILE: 98PL117
SCALE: 10:1
SHEET: 1 OF 1

Transformer:

TC16-161TX+
Upgraded Version*

Surface Mount
RF Transformer

TC16-161T+

50Ω 0.6 to 160 MHz

CASE STYLE: AT224-1A
PRICE: $2.59 ea. QTY (20)
$1.59 ea. QTY (100)

*Addition of Top hat™ feature
Benefits
• Allows faster pick-and-place
• Enables visual identification marking

+RoHS Compliant
The +Suffix identifies RoHS Compliance. See our web site for RoHS Compliance methodologies and qualifications

Maximum Ratings

Operating Temperature	-40°C to 85°C
Storage Temperature	-55°C to 100°C
RF Power	0.25 W
DC current	30mA

Permanent damage may occur if any of these limits are exceeded.

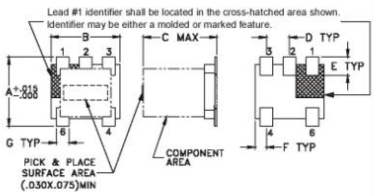
Pin Connections

PRIMARY DOT	6
PRIMARY	4
SECONDARY DOT	1
SECONDARY CT	2
SECONDARY	3
NOT USED	5

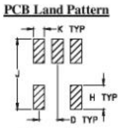
Features
• plastic base with solder plated leads
• excellent amplitude unbalance, 0.1dB typ. and phase unbalance, 0.5 deg. typ.

Applications
• impedance matching
• balanced amplifier

Outline Drawing AT224-1A



PCB Land Pattern



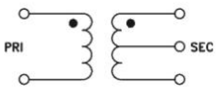
Suggested Layout,
Tolerance to be within ±.002

Transformer Electrical Specifications

Ω RATIO (Secondary/Primary)	FREQUENCY (MHz)	INSERTION LOSS*		
		3 dB MHz	2 dB MHz	1 dB MHz
16	0.6-160	0.6-160	1.5-120	3-80

* Insertion Loss is referenced to mid-band loss, 0.6 dB typ.

Config. A



Available Tape and Reel at no extra cost

Reel Size	Devices/Reel
7"	20, 50, 100, 200, 500
13"	1000, 2000

Outline Dimensions (inch/mm)

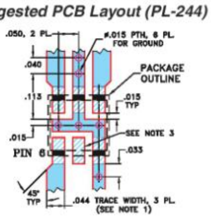
A	B	C	D	E	F
.150	.150	.160	.050	.040	.025
3.81	3.81	4.06	1.27	1.02	0.64

G	H	J	K	wt
.028	.065	.190	.030	grams
0.71	1.65	4.83	0.76	0.15

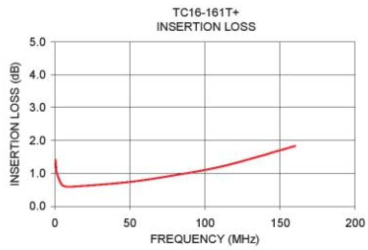
Typical Performance Data

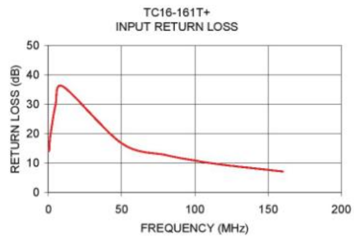
FREQUENCY (MHz)	INSERTION LOSS (dB)	INPUT R. LOSS (dB)
0.60	1.40	14.15
1.00	1.16	17.14
1.50	0.99	19.51
5.00	0.64	29.93
10.00	0.59	36.06
50.00	0.74	16.77
80.00	0.94	12.74
100.00	1.10	10.87
120.00	1.31	9.38
160.00	1.83	7.18

Demo Board MCL P/N: TB-145
Suggested PCB Layout (PL-244)



NOTES: 1. TRACE WIDTH IS SHOWN FOR ROGERS RO4350B WITH DIELECTRIC THICKNESS .020" ± .0015"; COPPER: 1/2 OZ. ON EACH SIDE. FOR OTHER MATERIALS TRACE WIDTH MAY NEED TO BE MODIFIED.
2. BOTTOM SIDE OF THE PCB IS CONTINUOUS GROUND PLANE.
3. THIS PAD IS NOT REQUIRED FOR AT224 CASE STYLE.
DENOTES PCB COPPER LAYOUT WITH SMOBC (SOLDER MASK OVER BARE COPPER)
DENOTES COPPER LAND PATTERN FREE OF SOLDER MASK





Notes
A. Performance and quality attributes and conditions not expressly stated in this specification document are intended to be excluded and do not form a part of this specification document.
B. Electrical specifications and performance data contained in this specification document are based on Mini-Circuit's applicable established test performance criteria and measurement instructions.
C. The parts covered by this specification document are subject to Mini-Circuits standard limited warranty and terms and conditions (collectively, "Standard Terms"); Purchasers of this part are entitled to the rights and benefits contained therein. For a full statement of the Standard Terms and the exclusive rights and remedies thereunder, please visit Mini-Circuits' website at www.minicircuits.com/MCLStore/terms.jsp

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Variable Capacitor:

SPRAGUE GOODMAN

FILMTRIM® PLASTIC DIELECTRIC CAPACITORS SG-402H

10 mm TOP / BOTTOM & SIDE ADJUST

SPECIFICATIONS

Voltage Rating: 200 VDC (High temp PTFE), 100 VDC (all others)

Dielectric Withstanding Voltage: 300 VDC (High temp PTFE), 200 VDC (all others)

Contact Resistance: 0.010 Ohms max

Insulation Resistance: 10^4 megohms min

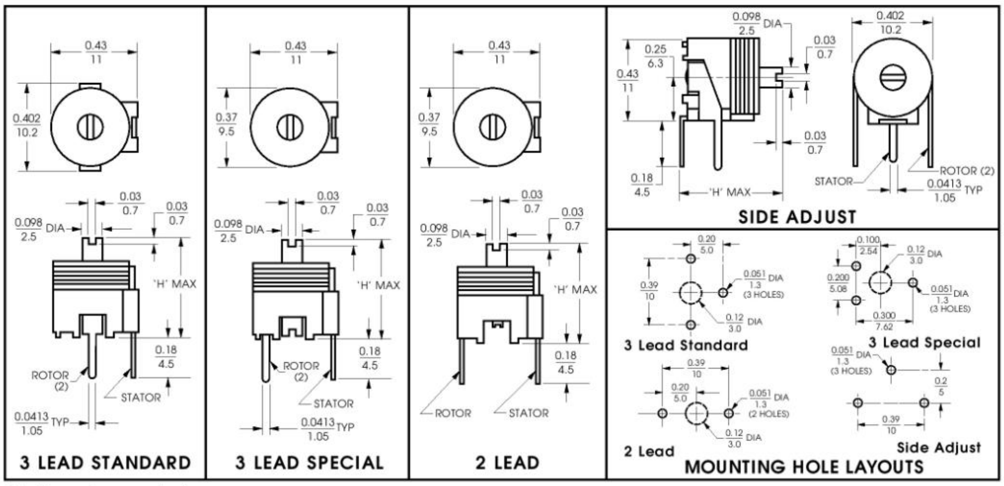
Torque: 15 to 360 g-cm (0.2 to 5 oz-in)

FEATURES

- Choice of dielectrics: High Temp PTFE, Standard PTFE, Polypropylene (PP), or Polycarbonate (PC)
- Linear capacitance change vs. rotation
- Wide capacitance ranges

Dielectric	Capacitance (pF) min	Capacitance (pF) max	Q min (1MHz)	TCC (ppm/°C)	Operating Temperature (°C)	'H' max in/mm	Color Code	Top/Bottom 3 Lead-Std. Model No.	Top/Bottom 3 Lead-Spec. Model No.	Top/Bottom 2 Leads Model No.	Side Model No.
PTFE*, High Temp	2.5	15.0	1500	0 ± 250	–40 to +125	0.402/10.2	Red	GXF15000	GXF15003	GXF15004	GXT15000
	3.0	25.0	1500	0 ± 250	–40 to +125	0.402/10.2	Clear	GXF25000	GXF25003	GXF25004	GXT25000
	4.0	40.0	1500	0 ± 250	–40 to +125	0.402/10.2	Yellow	GXF40000	GXF40003	GXF40004	GXT40000
	5.5	60.0	1500	0 ± 250	–40 to +125	0.449/11.4	Blue	GXF60000	GXF60003	GXF60004	GXT60000
	6.0	75.0	1500	0 ± 250	–40 to +125	0.449/11.4	Violet	GXF75000	GXF75003	GXF75004	GXT75000
	8.0	90.0	1500	0 ± 250	–40 to +125	0.488/12.4	Orange	GXF90000	GXF90003	GXF90004	GXT90000
PTFE	2.0	13.0	1500	0 ± 400	–40 to +85	0.402/10.2	Blue	GXC13000	GXC13003	GXC13004	GXD13000
	3.0	26.0	1500	0 ± 350	–40 to +85	0.402/10.2	Green	GXC26000	GXC26003	GXC26004	GXD26000
	3.5	38.0	1500	0 ± 300	–40 to +85	0.402/10.2	Clear	GXC38000	GXC38003	GXC38004	GXD38000
	6.0	60.0	1500	0 ± 300	–40 to +85	0.449/11.4	Yellow	GXC60000	GXC60003	GXC60004	GXD60000
	7.0	75.0	1500	0 ± 300	–40 to +85	0.449/11.4	Red	GXC75000	GXC75003	GXC75004	GXD75000
	8.0	90.0	1500	0 ± 300	–40 to +85	0.488/12.4	Violet	GXC90000	GXC90003	GXC90004	GXD90000
	10.0	150.0	1500	0 ± 300	–40 to +85	0.488/12.4	Orange	GXC15100	GXC15103	GXC15104	N/A
PP	2.0	15.0	1000	0 ± 400	–40 to +70	0.402/10.2	Blue	GYC15000	GYC15003	GYC15004	GYD15000
	3.0	20.0	1000	0 ± 400	–40 to +70	0.402/10.2	Green	GYC20000	GYC20003	GYC20004	GYD20000
	3.5	40.0	1000	0 ± 350	–40 to +70	0.402/10.2	Clear	GYC40000	GYC40003	GYC40004	GYD40000
	4.5	65.0	1000	0 ± 350	–40 to +70	0.402/10.2	Yellow	GYC65000	GYC65003	GYC65004	GYD65000
PC	8.0	80.0	200	0 ± 200	–40 to +85	0.402/10.2	Red	GZC80000	GZC80003	GZC80004	GZD80000
	9.0	100.0	200	0 ± 400	–40 to +85	0.449/11.4	Violet	GZC10100	GZC10103	GZC10104	GZD10100
	9.0	120.0	200	0 ± 350	–40 to +85	0.449/11.4	Orange	GZC12100	GZC12103	GZC12104	GZD12100
	10.0	150.0	200	0 ± 350	–40 to +85	0.472/12.0	Orange	GZC15100	GZC15103	GZC15104	GZD15100
	12.0	180.0	200	0 ± 350	–40 to +85	0.472/12.0	Orange	GZC18100	GZC18103	GZC18104	GZD18100

* Gold plated metal parts are standard on GXF and GXT models shown above.



All dimensions are in / mm.

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