Design of a Temperature Sensitive Band Pass Filter for a Chipless Harmonic Radar Sensor

Jordi Romeu Gómez

Masterarbeit (D-2196-M)
General Information

Type of Thesis: Masterarbeit
Thesis Title: Design of a Temperature Sensitive Band Pass Filter for a Chipless Harmonic Radar Sensor
Thesis Number: D-2196-M

Department: Institute for Microwave Engineering and Photonics
Mikrowellentechnik – Microwave Engineering
Examiner: Prof. Dr.-Ing. Rolf Jakoby

Author: Jordi Romeu Gómez
Supervisor: Dipl.-Ing. B.Kubina
Dipl.-Ing. C.Mandel

Start Date: 14.01.2013
Date of Submission: 14.07.2013
Examination Date: 25.07.2013

Affidavit/Eidesstattliche Versicherung

Rechtlich bindend ist nur der deutsche Text unten. Die englische Übersetzung dient nur dem besseren Verständnis.

Hereby I assure that I made the thesis by myself without third parties’ help only with the specified sources and aids. All figures, which were gathered from the sources, were marked as such. This thesis, in same or similar form, has not been available to any audit authority yet.

Hiermit versichere ich, die vorliegende Arbeit ohne Hilfe Dritter nur mit den angegebenen Quellen und Hilfsmitteln angefertigt zu haben. Alle Stellen, die aus den Quellen entnommen wurden sind als solche kenntlich gemacht worden. Diese Arbeit hat in gleicher oder ähnlicher Form noch keiner Prüfungsbehörde vorgelegen.

_________________________  ________________
Place, Date Author
Passive chipless wireless sensor tags are used in environments with specific conditions where chips, wires or power supplies cannot be used. Nowadays a huge part of these sensors use Surface Acoustic Wave (SAW) technology to encode the information and backscatter based in time-domain. However, this technology is limited by acoustic wave's propagation properties. Furthermore non-SAW chipless sensors have been presented. But these classical chipless systems are affected by radio clutter problems. In this thesis a novel passive chipless temperature sensor is presented which uses harmonic radar principles. It codifies the temperature as a spectral signature in frequency domain is presented. The tag uses the first harmonic frequency of the transmitted wave of the reading system to send the information to the reading system. In this case a frequency doubler is used to generate the first harmonic backscatter signal. Also the tag consists of a temperature sensitive band pass filter that uses a Dielectric Resonator (DR) to codify the temperature in frequency domain before the frequency doubler. In addition, the tag has one receiver antenna at the fundamental frequency to receive the incident wave of the reading system and one transmitter antenna at the first harmonic to send the harmonic radar backscatter signal. Both antennas are designed as stacked patch antennas because of their design simplicity and good performance. Wireless measurements were performed which proof the sensor concept. The reading was done with a distance range of up to 2.8 m and a temperature range between 20 °C and 100 °C.
# Contents

1 Introduction ................................. 4

2 Harmonic Radar Temperature Sensing ............................... 5

3 Temperature Sensitive Band Pass Filter .............................. 7
   3.1 Dielectric Resonator (DR) ............................................. 7
      3.1.1 Properties .......................................................... 7
      3.1.2 Cylindrical DR ..................................................... 8
         3.1.2.1 Resonant Frequency ......................................... 8
         3.1.2.2 Q Factor ....................................................... 11
      3.2 Band Pass Filter design .......................................... 12
         3.2.1 Desired performance .......................................... 12
         3.2.2 Filter topology ................................................ 13
         3.2.3 Principal parameters ........................................ 14
      3.3 Materials ............................................................. 15
      3.4 Simulation ........................................................... 16
      3.5 Final design ........................................................ 20
         3.5.1 Simulated Results .............................................. 20
         3.5.2 Measured Results .............................................. 21

4 Antennas .................................................. 24
   4.1 Patch antenna ...................................................... 24
      4.1.1 Design ............................................................. 25
   4.2 Bandwidth improvements ........................................... 26
      4.2.1 NEGCOMA and REGCOMA ...................................... 27
      4.2.2 Stacked patch antennas ....................................... 27
   4.3 Receiver antenna .................................................... 29
   4.4 Transmitter antenna ................................................ 31

5 Frequency doubler ............................................... 34
   5.1 Zero-bias Schottky diode .......................................... 34
   5.2 Matching Networks .................................................. 35
   5.3 Measurements ....................................................... 37

6 Measurement of the tag ............................................. 40
   6.1 Wired measurement ................................................ 40
   6.2 Wireless measurement ............................................... 42
      6.2.1 Unshielded tag ................................................ 42
      6.2.2 Shielded Tag ................................................... 46

7 Conclusion ...................................................... 51

A Layouts ......................................................... 52
1 Introduction

Nowadays passive wireless sensors are growing up in industrial processes. They can be used in locations with difficult access. Also they are useful in situations where no power supply can be used. Moreover chipless passive sensors can be used in situations where chip-based systems cannot, as for example in high temperature environments.

A huge part of the passive chipless sensors use Surface Acoustic Wave technology to encode the information and backscatter with time-domain encoding. This technology is limited by acoustic wave's propagation properties. To solve this limitation some electromagnetic approaches of a time-signature-based wireless build up with ID which allows the integration of several sensors have been presented [1].

The aim of this thesis is to present a passive chipless harmonic radar temperature sensor which allows to codify the information in frequency-domain instead of the time-domain. This thesis proves the concept of codifying the sensed information in frequency. Since the tag information is backscattered in another frequency band than the interrogation signal of the reader, many problems of chipless systems are reduced. The radar clutter is minimized and reference measurements are not necessary.

To achieve that aim, a harmonic radar temperature sensor has been designed and a prototype has been fabricated. The reader scans at a fundamental frequency range. The tag receives these frequencies with a receiver antenna. For this a antenna stacked patch is used. The received signal goes to a temperature sensitive band pass filter that codifies the temperature in frequency domain. This filter uses a Dielectric Resonator (DR) to shift the central frequency of the band pass with the temperature. A frequency doubler generates the first harmonic of the filtered frequency. This first harmonic is backscattered to the reader with another stacked patch antenna.

The wireless measurement results of this sensor are presented in this thesis in chapter 6. The temperature sensitive band pass filter is presented in chapter 3. A short introduction of the harmonic sensing principle is presented in chapter 2. The harmonic generator such as frequency doubler is presented in chapter 5. And the stacked patch antennas are presented in chapter 4.
2 Harmonic Radar Temperature Sensing

The aim of this thesis is to prove that a passive chipless tag can codify information in frequency domain and sends it at harmonic frequency to the reading system, which has first sent the fundamental frequency to the tag.

Harmonic radar sensing is based on the idea that the tag receives the input signal at the fundamental frequency $f_0$ and it transmit the information in harmonic frequency

$$f_{\text{harm}} = n \cdot f_0 \quad n = 2, 3, 4 \ldots$$

(2.1)

This idea seeks to avoid some traditional problems of radar sensing as the reflections or the need of reference measurement.

When the reading system scans the room it sends some tones inside a bandwidth. In traditional backscattering systems the reading system read at the same frequencies that it sends. Thus it reads reflections of its own signal, that makes hard the sensing. Also the own tag's antennas send a reflection of this signals, so it is needed to have reference measurements to be able to sense. However, if the tag uses another frequency to send back the information to the reading system these reflections are avoided as in figure 2.1 shown.

![Figure 2.1: a) Traditional backscattering sensing. b) Harmonic sensing](image)

Since the objective is also to codify the information (temperature) in frequency instead of amplitude a narrow band in the backscatter signal is desired. Thus, the narrower is the band the more frequencies could be use to detect the temperature and so the system could have a larger range of detecting temperatures or greater sensitivity. The part of the system which allows that is the passband filter which should change its central bandpass with temperature.
Passive tag means no power supply, so the transmitted power of the reading system is the only source of energy of the system. Thus the losses in the system should be minimized to be sure that the transmitted information of the tag could be read by the reading system. Then the antennas of the tag should have a great gain and the filter should have low losses in the pass band. Also the same is applied to the harmonic generator. It should have a low conversion loss or large conversion gain so most of the received power of the reading system could be used to send back the information. The passive sensors can be used in places of difficult access or with problems to put a wire. The whole system is represented in figure 2.2.

![Diagram of harmonic radar temperature sensor](image)

Figure 2.2: Schematic harmonic radar temperature sensor.

Chipless because usually chips do not support well higher temperatures so it is desired that the tag could sense different ranges of temperatures as for example parts of machines with hundreds of degrees.

The whole design is planar because a planar device can be placed easily and also they are cheap and easy to fabricate.

Finally harmonic radar temperature sensing means that the tag codifies the temperature in harmonic frequency and sends it back to the reading system.
3 Temperature Sensitive Band Pass Filter

An important element of the tag is the band pass filter. It must shift the central band pass frequency with the change of temperature and should have a narrow bandwidth. A simple way to do this is using some material that changes its frequency's properties with temperature. In this project is used a Dielectric Resonator (DR) to achieve this aim.

3.1 Dielectric Resonator (DR)

A DR is a piece of dielectric with free space boundaries and high $\varepsilon_r$ that resonates at certain frequencies. The frequency's value depends on the physical geometry of the piece, size, $\varepsilon_r$ of the material and the mode. Dielectric resonators are used in many fields of microwaves for their great properties. They can be used as antenna, oscillators or filter and reduce the size of them. They can work at elevate frequencies with low losses. In this project is used as band pass filter.

3.1.1 Properties

Some of principal properties have been mentioned before and most are mentioned in [2]:

- DRs usually have a high $\varepsilon_r$, usually between 10 and 100. Since the size of the DR depends strongly on $\frac{\lambda_0}{\varepsilon_r}$, it is significantly reduced and the system become more compact and lighter.
- They are made of low loss materials with $\tan \delta$ over $10^{-4}$ or less. So the losses are reduced.
- DRs offer simple coupling schemes to almost all transmission lines used in microwave technology.
- The inherent conductor loss is not present in DRs. This is really attractive for applications as millimetre wave antennas, where the metal loss can be high.
- Each mode in a DR has an unique internal/external field distribution.
- The operating bandwidth can be selected over a wide range by changing the right parameters of the resonator.
- Relays on the application, the unloaded Q-Factor ($Q_u$) can be strongly high or low. For filters or oscillators applications $Q_u$ is normally inversely proportionate to $\tan \delta$ and it fast goes to $10^4$. On the other hand, when a DR is used as antenna its $Q_u$ is reduced significantly between 10 or 100, because in this case the aim is to radiate power in a greater bandwidth.

This last property can be explained for the definition of Q-factor showed in [3]. Q-factor is the ratio between the maximum energy stored in the resonator ($W_{\text{max}}$) and the average power dissipated ($P_d$) in the resonator at resonance frequency ($\omega_0$)

$$Q = \frac{\omega_0 \cdot W_{\text{max}}}{P_d},$$

This last property can be explained for the definition of Q-factor showed in [3]. Q-factor is the ratio between the maximum energy stored in the resonator ($W_{\text{max}}$) and the average power dissipated ($P_d$) in the resonator at resonance frequency ($\omega_0$)

$$Q = \frac{\omega_0 \cdot W_{\text{max}}}{P_d},$$

but $P_d$ is formed for radiated power and loss power

$$P_d = P_{\text{rad}} + P_{\text{loss}},$$
so from (3.1) and (3.2) comes

\[
\frac{1}{Q_u} = \frac{P_{rad}}{\omega_0 \cdot W_{max}} + \frac{P_{loss}}{w_0 \cdot W_{max}} = \frac{1}{Q_{rad}} + \frac{1}{Q_{loss}}
\]

(3.3)

and the approximation

\[
Q_{loss} \approx \frac{1}{\tan \delta}.
\]

(3.4)

Equation (3.3) shows that the relation between \(Q_u\) and \(Q_{rad}\) and \(Q_{loss}\) is like a parallel resistance circuit. Also that mean that the lower Q-factor dominates the unloaded Q-factor and it is the maximum Q-Factor that could be achieved. Thus, it is possible to use DRs as filters, oscillators or antennas.

It must be noted that \(Q_{rad}\) variates with modes, shape and size. Also if the system is shielded, there would be no radiated power. As a result, with the same material or DR it is possible to get the different Q-Factors.

3.1.2 Cylindrical DR

The cylindrical DR is one of the most popular studied types. It has been studied as filter, oscillator or antenna. However, in all the applications it is important to understand the modes and their resonant frequency.

3.1.2.1 Resonant Frequency

There are different ways to solve the fields in a DR. One of the most simple is the magnetic wall model [4], which gives typically smaller values than experimental ones, about 10 percent. This method consists in to consider that the DR is in a contiguous magnetic-wall waveguide. With that, the problem becomes a waveguide problem.

In [5] Itoh and Rudokas describe another method that allows to have more precision in the resonant value. In this method, a DR with \(\varepsilon_{DR}\), radius \(a\) and height \(h_{DR}\) is placed on a dielectric substrate with \(\varepsilon_{sub}\) and thickness \(t\) as shows figure 3.1. The \(\varepsilon_{DR}\) is really higher than \(\varepsilon_{sub}\). To simplify the analysis microstrip lines are not taken into account.

![Figure 3.1: DR placed on substrate.](image)
Then, with a simplification is possible to solve the eigenvalue equation. The basic idea comes from the observation that in resonators with high Q most of the energy keeps inside of the DR, zone 1, and the fields decrease exponentially in regions 2, 3 and 4. It takes into account that the regions 5 and 6 have less power than the others regions, so it is possible to ignore them an get small error. With that situation, it is just necessary to match the field in the boundaries surfaces. For the TE modes in each zone:

\[
\begin{align*}
H_{z1} &= H_{01} \cdot \sin(\beta \cdot (z - z_0)) \cdot J_0(h \cdot r) \\
H_{z2} &= H_{02} \cdot \sin(\beta \cdot (z - z_0)) \cdot K_0(p \cdot r) \\
H_{z3} &= H_{03} \cdot e^{-\gamma(z-d)} \cdot J_0(h \cdot r) \\
H_{z4} &= H_{04} \cdot \text{sh}(\xi \cdot (z - z_0)) \cdot J_0(h \cdot r)
\end{align*}
\]

where,

\[
\begin{align*}
\beta^2 &= \varepsilon_{\text{DR}} \cdot k_0^2 - h^2 = k_0^2 + p^2 \\
p^2 &= (\varepsilon_{\text{DR}} - 1) \cdot k_0^2 - h^2 \\
\gamma^2 &= h^2 - k_0^2 = (\varepsilon_{\text{DR}} - 1)k_0^2 - \beta^2 \\
\xi^2 &= h^2 - \varepsilon_{\text{sub}} \cdot k_0^2 = (\varepsilon_{\text{DR}} - \varepsilon_{\text{sub}})k_0^2 - \beta^2 \\
k_0 &= \omega_0 \cdot \sqrt{\varepsilon_0 \cdot \mu_0} = \frac{2 \cdot \pi \cdot f_0}{c}
\end{align*}
\]

where \( J_0 \) and \( K_0 \) are Bessel function and modified Bessel function of order 0 respectively.

Due to the TE mode, the field components E\(z\), E\(r\) and H\(z\) are zero. Therefore, the continuity conditions are applied on components H\(z\) and E\(\theta\) at \( r = a, \ 0 < z < d \) and on E\(\theta\) and H\(r\) at \( z = 0 \) and \( d, 0 < r < a \). The eigenvalue system becomes:

\[
\frac{J'_0(h \cdot a)}{h \cdot J_0(h \cdot a)} + \frac{K'_0(p \cdot a)}{p \cdot K_0(p \cdot a)} = 0
\]

\[
\beta \cdot h_{\text{DR}} = q \cdot \pi + \tan^{-1} \left( \frac{\gamma}{\beta} \right) + \tan^{-1} \left( \frac{\xi}{\beta} \cdot \coth(\xi \cdot t) \right) , \ q = 0,1,2, \cdots
\]

The solution should satisfy (3.9), (3.14) and (3.15) simultaneously. Once \( k_0 \) is found, the resonant frequency \( f_0 \) could be obtained from (3.13), where \( c \) is light speed in vacuum.

The roots of (3.14) for a selected \( k_0 \) correspond to different radial fields distribution. They are ordered by index \( m \). Also the fields distribution order changes in axial direction. They have the index \( q \) as it shows in (3.15). Finally, the field in circumferential direction keeps constant, thus it takes the value 0. With all these considerations the modes are designated as \( \text{TE}_{0mq} \) and the lowest mode is \( \text{TE}_{010} \). This mode is also known as \( \text{TE}_{01\delta} \). From now, this method is referenced in this thesis as Itoh-Rudokas method.

This notation of the modes comes from the need to count in axial directions less than unit number of half-wavelength. That is indicated with the variable \( \delta \). The order variables \( n,m,q \) are integers. Also the notation is \( \text{TE}_{n,m,q+\delta} \). This notation is used in any kind of mode (TM or HEM, too). There are also other nomenclatures in literature [2], but this is the most common.

An improvement of Itoh-Rudokas method is presented in [6]. The presented method takes into account the effective relative dielectric constant \( \varepsilon_{\text{eff}} \) from the substrate where the DR is placed and included some closed design expressions for the mode \( \text{TE}_{015} \), usually the used mode. It also takes into account the distance between the DR and metallic planes on top and bottom of the system. As shown in figure 3.2, H\(2\) is the distance from the top of DR to the top metallic plane and H\(1\) is the distance from DR to the bottom metallic plane.

3.1 Dielectric Resonator (DR)
With these considerations, the eigenvalue equation for radial wavenumber \( h \) becomes

\[
\frac{J_0(h \cdot a)}{h \cdot J_0(h \cdot a)} + \frac{K_0(p \cdot a)}{p \cdot K_0(p \cdot a)} = 0
\]  

(3.16)

where \( p \) is

\[
p = \sqrt{(\epsilon_{\text{eff}} - 1) \cdot k_0^2 - h^2}
\]  

(3.17)

and \( \epsilon_{\text{eff}} \) is approximated by

\[
\epsilon_{\text{eff}} = \frac{\epsilon_{\text{eff}_a} + \epsilon_{\text{eff}_b}}{2}
\]  

(3.18)

where

\[
\epsilon_{\text{eff}_a} = \epsilon_{\text{DR}} - (\epsilon_{\text{DR}} - \epsilon_{\text{eff}_0}) \cdot \frac{H_1}{a}, \quad \frac{H_1}{a} \leq 1
\]  

(3.19)

\[
= \epsilon_{\text{eff}_0}, \quad \frac{H_1}{a} > 1
\]  

(3.20)

\[
\epsilon_{\text{eff}_b} = \epsilon_{\text{DR}} - (\epsilon_{\text{DR}} - \epsilon_{\text{eff}_0}) \cdot \frac{H_2}{a}, \quad \frac{H_1}{a} \leq 1
\]  

(3.21)

\[
= \epsilon_{\text{eff}_0}, \quad \frac{H_1}{a} > 1
\]  

(3.22)

finally,

\[
\epsilon_{\text{eff}_0} = \frac{h_0^2}{k_0^2}
\]  

(3.23)

\( h_0 \) is the radial wavenumber that should have an infinitely long cylinder with the same radius \( a \). In other words, case (3.14) with \( p \) as (3.10).

Equation (3.16) can be approximated for the mode TE\(_{015}\) to a closed form. It is useful to design DR’s dimensions or calculate the resonant frequency. The expression are for \( h \) and \( h_0 \)

\[
h \cdot a = 0.951 \cdot p_{01} + 0.222 \cdot \sqrt{(\epsilon_{\text{eff}} - 1) \cdot (k_0 \cdot a)^2 - 0.951 \cdot p_{01}^2}
\]  

(3.24)

\[
h_0 \cdot a = 0.951 \cdot p_{01} + 0.222 \cdot \sqrt{(\epsilon_{\text{DR}} - 1) \cdot (k_0 \cdot a)^2 - 0.951 \cdot p_{01}^2}
\]  

(3.25)

where \( p_{01} = 2.405 \) is the first root of \( J_0(x) = 0 \).

The eigenvalue equation for the axial wavenumber \( \beta \) is

\[
\beta \cdot h_{\text{DR}} = q \cdot \pi + \tan^{-1}\left(\frac{\gamma}{\beta} \cdot \coth(\gamma \cdot H_2)\right) + \tan^{-1}\left(\frac{\xi}{\beta} \cdot \coth(\xi \cdot H_1)\right) \quad q = 0,1,2,\cdots
\]  

(3.26)

Where \( \gamma \) and \( \xi \) are defined as (3.11) and (3.12). When both distances, \( H_1 \) and \( H_2 \), tend to infinity \( \epsilon_{\text{eff}} \) becomes \( \frac{h_0^2}{k_0^2} \) so it becomes the expression used for isolated resonator.

3.1 Dielectric Resonator (DR)
As in the Itoh-Rudokas method, the solution should satisfy (3.9), (3.26) and one of (3.16) or (3.24).

This second presented method is referenced from now as Mongia method when (3.16) is used and Approximation Mongia method when (3.24) is used.

### 3.1.2.2 Q Factor

As is explained in 3.1.1, Q-Factor has a strong dependence on material, mode, shape and size. Furthermore, the Q-Factor could be separated in lossy Q-Factor and radiated Q-Factor. The first one is based on the material losses and it could be approximated by equation (3.4). While the second one hangs on form, mode, material and size.

For cylindrical DR’s $Q_{\text{rad}}$ depends on material’s $\epsilon_r$, the mode and the aspect ratio, which is defined as $\frac{h_{\text{DR}}}{a}$. The aspect ratio is an important parameter. This ratio allows to use some closed formula in determinate range to get different approximation values for parameters as $Q_{\text{rad}}$ or resonant frequency [2]. However, the formulas, that are given, are for an isolated DR. That means, that for each mode in an isolated DR exists equivalent mode for a DR placed on a metal ground, because the structure acts like an isolated DR with double height for modes with electric wall in its plane of symmetry. This equivalence are in table 3.1.

<table>
<thead>
<tr>
<th>isolated DR</th>
<th>DR metal plane</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\text{TE}_{011+\delta}$</td>
<td>$\text{TE}_{01\delta}$</td>
</tr>
<tr>
<td>$\text{HE}_{11+\delta}$</td>
<td>$\text{HE}_{11\delta}$</td>
</tr>
<tr>
<td>$\text{EH}_{111+\delta}$</td>
<td>$\text{EH}_{11\delta}$</td>
</tr>
<tr>
<td>$\text{TM}_{01\delta}$</td>
<td>$\text{TM}_{01\delta}$</td>
</tr>
<tr>
<td>$\text{HE}_{21\delta}$</td>
<td>$\text{HE}_{21\delta}$</td>
</tr>
<tr>
<td>$\text{EH}_{21\delta}$</td>
<td>$\text{EH}_{211+\delta}$</td>
</tr>
</tbody>
</table>

Also, for the mode $\text{TE}_{01\delta}$ is used the formula in isolated DR of the mode $\text{TE}_{011+\delta}$, which is valid for $0.2 \leq \frac{h_{\text{DR}}}{a} \leq 2$.

$$Q_{\text{rad}} = 0.03628 \cdot \epsilon_{\text{r}_{\text{DR}}}^{2.38} \cdot \left[ -1 + 7.81 \cdot \left( \frac{h_{\text{DR}}}{a} \right) - 5.858 \cdot \left( \frac{h_{\text{DR}}}{a} \right)^2 + 1.277 \cdot \left( \frac{h_{\text{DR}}}{a} \right)^2 \right]$$  \hspace{1cm} (3.27)

Equation (3.27) is plotted in figure 3.3 for $\epsilon_{\text{r}_{\text{DR}}} = 1$ in the range set before. Thus, it is possible to get the $Q_{\text{rad}}$ for a specific DR multiplying this figure by $\epsilon_{\text{r}_{\text{DR}}}^{2.38}$, which is constant. For example, if $\epsilon_{\text{r}_{\text{DR}}} = 20$, then $Q_{\text{rad}_{\text{max}}} = 107.55$.

Figure 3.3 shows a maximum near aspect ratio as 1. In this case, an isolated DR with low losses would have a low radiation bandwidth and its Q-Factor would be the highest for the mode $\text{TE}_{011+\delta}$. Meanwhile, for aspect ratio near 0.2 its Q-Factor would be the lowest and it would have a the greatest radiation bandwidth. So the same happens for the mode $\text{TE}_{01\delta}$ for a cylindrical DR placed over of metal plane.
3.2 Band Pass Filter design

The band pass filter has a great influence on the system. Its performance conditions the design of the other parts of the tag.

3.2.1 Desired performance

The ideal filter should have a narrow bandwidth and it should have a high attenuation outside the pass band. That would make easier to detect the resonant frequency. Thus mean that the Q-Factor should be high.

An other important fact is how should be the frequency increase $\Delta f$ with temperature increase $\Delta T$. There is a compromise between sensitivity, temperature range and bandwidth. A low $\Delta f$ value for a fix $\Delta T$ would allow a higher temperature sensing with the same bandwidth, but it would have less sensitivity. That mean, the minimum value of $\Delta T$, that the system would detect, must be higher. In other hand, a high $\Delta f$ would detect smaller $\Delta T$. However, the maximum temperature value would be smaller too for the same bandwidth. These situations are represented in figure 3.4, where $f_0$ is the reference frequency and $f_1$ is the frequency for a fix $\Delta T$. Both are represented in axis with normalized frequency $F = \frac{f}{f_0}$.

![Figure 3.4: Different cases of $\Delta f$ with the same $\Delta T_0$.](image)

If the upper frequency of the bandwidth is in terms of $F$ 1.5, for the case A the maximum value of $\Delta T$, that the system would measure, would be 5 times $\Delta T_0$. Meanwhile, for the case B the maximum $\Delta T$ would be 50
However, case A would have better sensitivity than case B since it would shift faster the resonant frequency. So, the tag’s application will define which case is better. In this way, it is possible to define the parameter of absolute sensitivity \( S_{fT} \) as the ratio between \( \Delta f \) and \( \Delta T \) as in (3.28)

\[
S_{fT} = \frac{\Delta f}{\Delta T}.
\]  

(3.28)

With this parameter as reference, case A \( S_{fT_A} \) is 10 times case B \( S_{fT_B} \).

Finally, because it forms part of a RFID system it should introduce the minimum losses in band pass for all the central frequency sweep.

### 3.2.2 Filter topology

There are lots of different ways to use a DR to create the bandpass filter. The desired device must be planar. Thus, the coupling will be with microstrip lines. In [7] there are two possible configurations as shows figure 3.5 to bandpass filter design.

In figure 3.5 \( L \) represents the electrical length instead of the physical length. In type I, which is shown in figure 3.5(a), the resonant frequency of the filter is the DR’s itself for any \( L \). While in type II, figure 3.5(b), has a resonant frequency near to the DR’s on. In this case the resonant frequency of the filter depends on \( L \), but it is constant for a fix one [7].

For both topologies, when \( L \) is \( \frac{\lambda}{4} \) or an even multiple the S-parameter \( S_{21} \) is maximum [7]. This maximum value is function of coupling factors between the microstrip and the DR usually represented with \( \beta_i \). Strong values of \( \beta_i \) give higher \( S_{21} \). This parameter \( \beta_i \) has a dependence with the distance \( d_i \). Higher \( d_i \) has weak coupling and lower \( \beta_i \).

A way to simplify the systems is taking both distance equals. Then the system becomes symmetric and only depends on a coupling factor \( \beta \).

In this Thesis the used topology is the one from figure 3.5(a) with \( L = \frac{\lambda}{4} \) and \( d_1 = d_2 = d \). So, the systems resonant frequency could be easily obtained with the formulas in section 3.1.2.1 and keeping the losses as low as possible. Also the filter becomes a symmetric system.
3.2.3 Principal parameters

Once the topology and the performance has been defined it is possible to define some important parameters which have a great influence to achieve the aim.

For the temperature sensing is important to know how is the effect of rising the temperature in a DR. \(\tau_f\) is the coefficient that relates \(\Delta f\) with \(\Delta T\).

\[
\Delta f = f_0 \cdot \tau_f \cdot \Delta T \tag{3.29}
\]

then,

\[
f(\Delta T) = f_0 \cdot (1 + \tau_f \cdot \Delta T). \tag{3.30}
\]

This coefficient is formed from three other temperature coefficients: dielectric constant \(\tau_e\), cavity \(\tau_c\) and thermal expansion \(\alpha_L\). \(\tau_e\) is the most important coefficient. The others can be neglected with a low error in \(\tau_f\) so the approximation expression is:

\[
\tau_f = -\frac{\tau_e}{2} \tag{3.31}
\]

After placing the DR in the topology, the whole structure has new characteristics as a band pass bandwidth (BW) and a loaded Q-Factor \(Q_L\). Also the resonant frequency could change a bit. The bandwidth is delimited for the lower and upper frequency with \(-3\) dB from the resonant frequency in transmission \((S_{21})\). If \(f_1\) is the lower cut frequency and \(f_2\) the upper one, then:

\[
BW = f_2 - f_1. \tag{3.32}
\]

\(Q_L\) should to count all the losses in the system plus the \(Q_u\). The whole dissipated power of the system \(P_{\text{dis}}\) is formed by the power dissipated in the DR \(P_{\text{DR}}\) and the power dissipated in the rest of circuit called external power \(P_{\text{ext}}\). This \(P_{\text{ext}}\) defines an external Q-Factor \(Q_{\text{ext}}\). As in (3.1), the loaded Q-Factor is defined:

\[
Q_L = \frac{\omega_0 \cdot W_{\text{max}}}{P_{\text{dis}}} = \frac{\omega_0 \cdot W_{\text{max}}}{P_{\text{DR}} + P_{\text{ext}}} \tag{3.33}
\]

\(P_{\text{dis}}\) is the input power \((P_{\text{in}})\) that is not reflected \((P_{\text{ref}})\) or transmitted \((P_{\text{trans}})\). In this case, it is formed by the lost power and the radiated power.

\[
P_{\text{dis}} = P_{\text{in}} - P_{\text{ref}} - P_{\text{trans}} \tag{3.34}
\]

As in (3.3) \(Q_L\) can be expressed in function of \(Q_u\) and \(Q_{\text{ext}}\)

\[
\frac{1}{Q_L} = \frac{1}{Q_u} + \frac{1}{Q_{\text{ext}}} \tag{3.35}
\]

It is possible to define a coupling coefficient \(k\) as the ratio between the power dissipated in the resonator and the power dissipated in the rest of the structure [3].

\[
k = \frac{Q_u}{Q_{\text{ext}}} \tag{3.36}
\]

This coefficient shows how strong is the energy coupling. When \(k = 1\) it is said that the coupling is critical. That means that the resonator and the rest of the system dissipates the same power. Meanwhile, when \(k < 1\) is called undercritical and means that the resonator dissipates more power than the rest. Finally, when \(k > 1\) it is overcritical and the external losses dissipate more power than the resonator. As it is said in section 3.2.2, this coefficient is also known as \(\beta\) for the coupling between microstrip and DRs.
From (3.35) and (3.36) comes

\[ Q_u = (1 + k)Q_L \]  

(3.37)

This equation (3.37) shows that for critical coupling \( Q_L \) is the half of \( Q_u \).

Also, from (3.32) and \( f_0 \) is possible to define the loaded Q-Factor as:

\[ Q_L = \frac{f_0}{BW} \]  

(3.38)

which shows that if a low BW is desired then the \( Q_L \) should be high.

### 3.3 Materials

Since the aim of this project is to prove the possibility to codify information in frequency for a chipless RFID, the desired temperature range is low, it starts at 20°C and ends at 80°C. Also a high relation of \( \Delta f \) to the peak bandwidth (BW) is desired. Thus, it is used for the design a sample of material K–80 with a typical \( \varepsilon_r = 80 \pm 5\% \) at 9.4 GHz, \( \tan \delta < 10^{-3} \) and coefficient of dielectric constant \( \tau = -\frac{490 \cdot 10^{-6}}{\circ C} \). That mean a theoretical \( Q_{\text{loss}} = \frac{1}{\tan \delta} > 1000 \) and \( f_f = -\frac{490 \cdot 10^{-6}}{3} = 980 \cdot 10^{-6} \circ C^{-1} \). The sample is placed on a ROGERS RT 5880 substrate with thickness \( t = 1.575 \text{ mm} \), \( \varepsilon_{\text{sub}} = 2.2 \) and \( \tan \delta = 0.0009 \) at 10 GHz. Although the used substrate is with a thickness \( t = 1.575 \text{ mm} \), for the first simulations is used the same material with a thickness \( t = 0.8 \text{ mm} \).

The DR has a radius \( a = 12.7 \text{ mm} \) and a height \( h_{\text{DR}} = 10.4 \text{ mm} \). The ratio \( \frac{h_{\text{DR}}}{a} = 0.819 \) is near to the optimal value for a highest \( Q_{\text{rad}} \). In this case, it is about 2833, so from (3.3) \( Q_u > 1000 \). The resonant frequency for the mode \( \text{TE}_{01} \) is calculated using section 3.1.2.1 for the two methods and for the approximation from the second. The results are presented in table 3.2.

<table>
<thead>
<tr>
<th>Method</th>
<th>Method</th>
<th>( \varepsilon_{\text{DR}} = 80 )</th>
<th>( \varepsilon_{\text{DR}} = 85.3 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Itoh-Rudokas Method</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>(3.14),(3.15) and (3.13)</td>
<td>1.575</td>
<td>1.64056</td>
<td>1.575</td>
</tr>
<tr>
<td>Mongia Method</td>
<td>0.8</td>
<td>1.68664</td>
<td>0.8</td>
</tr>
<tr>
<td>(3.16),(3.26) and (3.13)</td>
<td>1.575</td>
<td>1.56493</td>
<td>1.575</td>
</tr>
<tr>
<td>Approximation Mongia Method</td>
<td>0.8</td>
<td>1.61542</td>
<td>0.8</td>
</tr>
<tr>
<td>(3.24),(3.26) and (3.13)</td>
<td>1.575</td>
<td>1.59842</td>
<td>1.575</td>
</tr>
<tr>
<td></td>
<td>0.8</td>
<td>1.64656</td>
<td>0.8</td>
</tr>
</tbody>
</table>

This sample is measured and it is found that it has a \( Q_u = 2590 \) and a \( \varepsilon_{\text{DR}} = 85.3 \). The resonant frequency's values are calculated again and also represented in table 3.2.

The decreasing of \( f_0 \) is almost the same for each method, nearly 51 MHz, for an increase of 5.3 in \( \varepsilon_{\text{DR}} \).

The system is design to have an impedance \( Z = 50 \Omega \), so the width of the microstrip in RT 5880 with \( t = 1.575 \text{ mm} \) around \( f = 1.56 \text{ GHz} \) is \( W = 4.83 \text{ mm} \). Also, \( \frac{a}{4} = 34.98 \text{ mm} \). And for a \( t = 0.8 \text{ mm} \) the values become \( f = 1.6 \text{ GHz} \), \( W = 2.2 \text{ mm} \) and \( \frac{a}{4} = 34.14 \text{ mm} \).


3.4 Simulation

In order to analyse the performance of the filter, EM full-wave simulations have been performed. These simulations investigate the input design parameters such as the distance $d_1 = d_2 = d$ on the unloaded Q-factor, $Q_L$ and transmission of the filter. It is used CST Microwave Studio and several simulation are run.

The system is defined as in figure 3.5(a). To avoid change the resonant frequency, the minimum value of $d$ should be greater than the radius of the DR $a$ plus the half of the microstrip width $W$. In this case $d \geq \frac{W}{2} + a$ mm. For the first simulations the substrate’s thickness is $t = 0.8$ mm. So $d \geq 13.8$ mm.

Simulations with different values of $d$ are run. Also three different cases are used. In the first one, an ideal DR is placed on an lossless substrate and perfect conducting metal. In the second case, the substrate and the metal are still ideal, meanwhile the DR has losses. Finally, in the third case, the simulation takes into account every loss from each material (DR, substrate and metal). The results are plotted in figure 3.6.

These results have a resonant frequency that shifts a little depending on the distance. For the first distance $d$, $f_0 = 1.601$ GHz. However, it is constant after $d = 18.7$ mm with $f_0 = 1.5998$ GHz. This difference in the smaller distance could be generated for the microstrip lines that are not taken into account in the theory.

This value of $f_0$ is near to the theoretical ones that are in table 3.2. In fact, in this case the error is less than 2.3% for all the methods and simulated values of $f_0$.

In addition, the results show two important characteristics. First, when $d$ increases then $S_{21}$ decreases. Thus, the transmitted power decreases, too. And the second one, there are nearly no differences between the second case and the third, which means that most of the losses of the filter are in the DR. These losses are around 5 dB at the resonant frequency and they keep constant with $d$.

Figure 3.6: S-parameter for DR placed on substrate where $d_1 = d_2 = d$ and $t = 0.8$ mm.

In the left, s-parameters from ideal case.
In the middle, s-parameters with only DR losses.
In the right, s-parameters with all losses.
To characterize better the system plots of the powers over $d$ are generated in figures 3.10 and 3.7 respectively. They also take into account the lossy cases as in figure 3.6.

In Figure 3.7 it is possible to see that when the distance $d$ increases the transmitted power from input to output $P_{\text{trans}}$ decreases and the reflected power from the input $P_{\text{ref}}$ increases, as is expected. However, the lost power $P_{\text{loss}}$ has more interest. In figure 3.7 all these power are normalized to the input power. For the case no material losses, its value is the radiated power of the system. Furthermore, $P_{\text{loss}}$ has the maximum value when $P_{\text{trans}}$ and $P_{\text{ref}}$ have similar values.

The DR is shielded to avoid radiation losses. The shield is simulated as a boundary with $E = 0$ at all directions.

In figure 3.8 as explained in section 3.1.2.1 the resonant frequency changes for the metallic plane in the normal direction of the DR. Another important difference from the unshielded DR is that the power of the interest mode TE$_{01\delta}$ decreases less with the increase of $d$. Also the second mode near 1.7 GHZ has more power than in figure 3.6.

In figure 3.9 the power over the distance is represented. The lost power $P_{\text{loss}}$ is equal to zeor for the ideal materials, which is reasonable since there are no radiated losses. For the cases DR losses and all materials losses the results are similar to figure 3.7 but with less value since the radiation component is zero. Also $P_{\text{trans}} + P_{\text{ref}}$ is higher than for the unshielded DR for all losses case.

To keep inside the design parameters it is also needed to know the effect of $d$ in $Q_L$ for both cases (shielded and unshielded DR). Figure 3.10 represents these curves.

What comes from these plots is that the shielded DR has a greater unloaded Q-Factor $Q_L$ than the unshielded one. Furthermore, $Q_L$ increases with $d$ until a value near the unloaded Q-Factor $Q_{Lu}$ for the shielded DR and ideal materials. However, when the losses are taken into account this difference is strongly reduced.

On the other side, the fabrication and measurement of the unshielded DR filter is easier. Also for sensing the temperature the DR should be directly in contact with the air. So the shield should have some holes that allow the air pass but not the electro-magnetic radiation under 2 GHz. Thus the final design is without shield.

3.4 Simulation
Figure 3.8: S-parameter for shielded DR placed on substrate where \( d_1 = d_2 = d \) and \( t = 0.8 \) mm. In the left, s-parameters from ideal case. In the middle, s-parameters with only DR losses. In the right, s-parameters with all losses.

Figure 3.9: Power for shielded DR placed on substrate where \( d_1 = d_2 = d \) and \( t = 0.8 \) mm. In the left, \( P_{\text{loss}} \) for the different losses cases. In the middle, \( P_{\text{trans}} \) for the different losses cases. In the right, \( P_{\text{ref}} \) for the different losses cases.

However the final design has a substrate RT 5880 with thickness \( t = 1.575 \) mm. For this reason the system is simulated again with this thickness. In this simulations only the all losses case is taken. In addition, it is important to note that now \( d \geq 15.115 \) mm.

3.4 Simulation
The first important difference, which shows figure 3.11, from the case \( t = 0.8 \) mm is that \( f_0 \) has changed to 1.558 GHz. This change is expected, as shown table 3.2 higher \( t \) has lower \( f_0 \). Also \( S_{21} \) decreases more slowly in relation to \( d \). This trend is shown in figure 3.12. Also this figure shows \( P_{\text{loss}} \) and \( P_{\text{ref}} \).

This trend to change more slowly appears also in \( Q_L \) in figure 3.13. To keep a high Q-factor is needed to lose transmitted power.

For this reason the range of \( d \) is limited to \( 20 \text{ mm} < d < 25 \text{ mm} \) in further simulations. In this range the Q-factor start to have values over 200 and \( P_{\text{trans}} \) is still over 20%.

In this range several simulation are made to see the effects of two different ways to fix the DR on the substrate: glue and tin. In this simulations some blocks with different size of each material are placed around the DR to fix it. Figure 3.14 shows changes in \( S_{21} \).

While for the glue there is nearly no differences in each size from the one without any fix, for the tin the resonant frequency becomes smaller when the size of the block increases. Thus, the glue is selected as the fixer material.

3.4 Simulation
Figure 3.12: Normalized Power over $d$ for $t = 1.575$ mm, unshielded case.

Figure 3.13: Loaded Q-Factor $Q_L$ for $t = 1.575$ mm, unshielded cas.

### 3.5 Final design

The final design is selected after several simulations that effect the principal parameters. It is important to note that the values are selected for having an optimal case. Thus it is possible that for other applications to change the threshold but for this case the values are taken to have a high $Q_L$ factor with a level of acceptable transmitted power.

#### 3.5.1 Simulated Results

The selected values are $f_0 = 1.558$ GHz, $d = 23.7$ mm, $Q_L = 601$ and $P_{\text{trans}} = 0.303$. All this values are obtained from a CST simulation with a mesh density of 4 cells per wavelength. To check that this results are quite accurate another simulation with higher mesh of 6.
Figure 3.14: $S_{21}$ for different sizes of glue and tin fix.

Figure 3.15: Simulated s-parameters for mesh 4 and 6.

Figure 3.15 shows that the accuracy of the previous values is good enough. $f_0$ changes not really much to a lower value of 1.556 GHz. That also happens to $Q_L$ which becomes 587. That means that $P_{\text{trans norm}}$ increases its value to 0.308.

### 3.5.2 Measured Results

With the results of the last simulation a prototype has been build. For the prototype three different fixers have been tested. Due to mechanical reasons the show different adhesion strengths. The best adhesion has
been achieved with the ROGERS 3001 Bonding film. They effects the electromagnetic charasterics, thus the resonant frequency changes at the same temperature from some measures to others.

The s-parameter at room’s temperature is measured. Figure 3.16 shows the results for two ways to fix the DR. The first one is without any kind of fixer. The DR is placed in the right point and the filter is measured. The second one cellophane tape is used to fix the DR at the right point.

![Figure 3.16: S\textsubscript{21} for the cellophane and no fixer at T = 22°C](image)

The shape of S\textsubscript{21} is really similar to the simulated one. Also the resonate frequency is nearly the same. The cellophane tape has a little influence in \( f_0 \) and the filter has less losses.

The filter is heated with a heating gun to rise its temperature. In this process are used three possible fixers. Two of them are the cellophane tape and the no fixer, which are described before. The third way is using the 3001 Bonding film provided from ROGERS, that allows the substrate to stick onto another substrate.

![Shift \( f_0 \) for film fixer](image)

![Shift \( f_0 \) for no fixer](image)

![Shift \( f_0 \) for cellophane tape fixer](image)

Figure 3.17: Shift of \( f_0 \) for different fixers.

As figure 3.17 shows, the shift of the resonant frequency is easily detectable. In the case of film fixer appears a valley between 40°C and 70°C. When the heating gun heats the filter the substrate starts to bend and changes the coupling between the DR and the microstrip lines. In other hand, in cellophane case the substrate
is fixed in other material to keep its form without bending. Also for the case of the film fixer the values of $f_0$, $Q_L$, and $P_{\text{loss}}$ are represented in table 3.3 for different temperatures.

Table 3.3: Measured value of resonant frequency for different values of T for the film fixer.

<table>
<thead>
<tr>
<th>T(°C)</th>
<th>$f_0$(GHz)</th>
<th>$Q_L$</th>
<th>$P_{\text{loss}}$(dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>22</td>
<td>1.5573</td>
<td>471.9</td>
<td>-5</td>
</tr>
<tr>
<td>32</td>
<td>1.5639</td>
<td>492.5</td>
<td>-4.6</td>
</tr>
<tr>
<td>39</td>
<td>1.5693</td>
<td>484.5</td>
<td>-5.1</td>
</tr>
<tr>
<td>52</td>
<td>1.574</td>
<td>526.2</td>
<td>-7.1</td>
</tr>
<tr>
<td>60</td>
<td>1.5791</td>
<td>656.3</td>
<td>-11.1</td>
</tr>
<tr>
<td>70</td>
<td>1.5841</td>
<td>560</td>
<td>-5.5</td>
</tr>
<tr>
<td>80</td>
<td>1.5899</td>
<td>548.5</td>
<td>-5.7</td>
</tr>
</tbody>
</table>

Finally figure 3.18 plots for the three fixer situations the value of the resonant frequency $f_0$ over the temperature. The three curves show that the relation is nearly lineal and it is nearly the same for the three cases.
4 Antennas

The tag should receive at $f_0$ and transmit at $2f_0$. Thus, the tag is formed by two antennas with different frequency range. To simplify the design both antennas are designed as patch antennas. The antenna which receives at $f_0$ is called in this thesis receiver antenna and the one which transmits at $2f_0$ is called transmitter antenna.

The receiver antenna should have a 3 dB-bandwidth at that covers at least a frequency range for temperature range from 20°C to 80°C. This range is calculated as

$$f_0(T = 20°C) = 1.556\, \text{GHz}$$
$$f_0(T = 80°C) = f_0(T = 20°C) \cdot (1 + \tau_f \cdot (80°C - 20°C)) = 1.595\, \text{GHz},$$

with

$$\tau_f = 410 \cdot 10^{-6} \, \text{°C}^{-1},$$

which is the maximum value from the datasheet of the DR's material.

However to avoid problems the antennas are designed to achieve a temperature range from 0°C to 100°C. That gives a lower frequency of $f_0(T = 0°C) = 1.544\, \text{GHz}$ and upper frequency $f_0(T = 100°C) = 1.608\, \text{GHz}$. Thus, the receiver antenna is design to achieve at least $f_l = 1.54\, \text{GHz}$ in the lower band and $f_u = 1.61\, \text{GHz}$ in the upper band. That makes a bandwidth at 3 dB

$$\text{BW}_{3\, \text{dB}} = 1.61\, \text{GHz} - 1.54\, \text{GHz} = 70\, \text{MHz}$$

Usually the BW is given in % or :1 defined as

$$\text{BW}(%) = \frac{2 \cdot (f_u - f_l)}{f_u + f_l} \cdot 100\% \quad \text{BW} \leq 100\%$$
$$\text{BW}(1) = \frac{f_u}{f_l} : 1 \quad \text{BW} \geq 100\%.$$

In this case the receiver antenna needs at less $\text{BW}(%) = 4.44\%$ or $\text{BW}(1) = 1.045 : 1$. This values are also for the transmitter antenna since $f_{u_{tx}} = 2 \cdot f_u$ and $f_{l_{tx}} = 2 \cdot f_l$. But $\text{BW}_{3\, \text{dB_{tx}}} = 2 \cdot \text{BW}_{3\, \text{dB}} = 140\, \text{MHz}$.

Compact, planar and metal mounting antennas with broadside radiation pattern are desired. Thus the receiver antenna and the transmitter antenna are designed as rectangular patch antenna. Also, to avoid mismatching problems, the substrate used is the same of the filter, ROGERS RT5880 with thickness $t = 1.575\, \text{mm}$.

4.1 Patch antenna

A patch antenna is a planar antenna which is typically formed for a really thin metal patch(thickness $t_{\text{patch}} << \lambda_0$) over a metal ground at height $h$ that is a small fraction of $\lambda_0$ [8]. This height is usually filled with dielectric substrate. It usually is designed to radiate in the normal direction of the patch. Also it is easy to design and allows different feeding systems as microstrip, coupled slot or coaxial. In figure 4.1 is represented a rectangular patch antenna with microstrip feeding.
On the other side, the substrate has a great impact on the performance of the patch antenna. A thick substrate with low \( \varepsilon_r \) has better efficiency and more bandwidth, which is usually less than a few percent. A thin substrate with high \( \varepsilon_r \) has tightly bound fields and reduces the size of the antenna, but it has higher losses, smaller bandwidth and less efficiency.

### 4.1.1 Design

For rectangular patches antennas the resonant frequency(\( f_r \)) of the dominant mode, that is TM\(_{010}\), depends on the length \( L \). It is usually near to \( \frac{\lambda}{2} \). To calculate the parameters of the antenna there are few models, the simple one but with good physical insight is the transmission line model [8].

This model considers the rectangular patch antenna as two slots antenna separated a distance \( L \) and with a size \( W \). Also it takes into account the fringing effects that appears for the finite dimensions \( L \) and \( W \) of the patch. The impact of the fringing effects is reduced when \( \frac{L}{h} \gg 1 \). However they influence the resonant frequency.

The steps to design the rectangular patch are:

1. Calculate \( W \) for an efficient performance with

\[
W = \frac{c}{2 \cdot f_r} \cdot \sqrt{\frac{2}{\varepsilon_r + 1}}. \tag{4.8}
\]

2. Determine \( \varepsilon_{\text{eff}} \) using

\[
\varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \cdot \frac{1}{\sqrt{1 + \frac{12 h}{W}}}. \tag{4.9}
\]

3. Approximate \( \Delta L \) with equation

\[
\Delta L = 0.412 \cdot h \cdot \frac{(\varepsilon_{\text{eff}} + 0.3)(\frac{W}{h} + 0.264)}{(\varepsilon_{\text{eff}} - 0.258)(\frac{W}{h} + 0.8)}. \tag{4.10}
\]
4. Finally calculate $L$ with

$$L = \frac{c}{2f_r \sqrt{\varepsilon_{eff}}} - 2\Delta L.$$  

(4.11)

Furthermore, it is possible to estimated the patch impedance $Z_{in}$ for the resonant frequency. The equivalent circuit for the patch antenna is two parallel admittance $Y_1 = G_1 + jB_1$ and $Y_2 = G_2 + jB_2$ that represent the two slots. Also $Y_1 = Y_2$, $G_1 = G_2$ and $B_1 = B_2$. This admittance at $f_r$ is separated nearly $\frac{\lambda}{2}$ in spite of the fringing effects. Then the transformed admittance of the second slot becomes

$$Y_2' = G_2' + jB_2' = G_1 - jB_1,$$  

(4.12)

so

$$Y_{in} = 2G_1,$$  

(4.13)

finally

$$Z_{in} = \frac{1}{Y_{in}} = \frac{1}{2G_1} = R_{in}$$  

(4.14)

and

$$G_1 = \frac{W}{120\lambda_0} \left( 1 - \frac{(k_0h)^2}{24} \right) \frac{h}{\lambda_{\text{ada}}} < \frac{1}{10}.$$  

(4.15)

Equation (4.14) does not take into account the mutual effects between the slots. These mutual effects are represented with $G_{12}$ and it is approximated with

$$G_{12} = \frac{1}{120\pi^2} \int_0^\pi \sin \left( \frac{k_0W}{2} \cos \Theta \right) \cos \Theta \cdot J_0(k_0L \sin \Theta) \cdot \sin^3 \Theta d\Theta$$  

(4.16)

then,

$$Z_{in} = \frac{1}{2(G_1 \pm G_{12}).}$$  

(4.17)

$Z_{in}$ is referenced to the slot 1. However it is possible to change this input impedance with an inset feed. This inset feed goes from the edge of slot 1 to the centre of the patch a distance $y_0$. This distance $y_0$ could be obtained from the approximation

$$Z_{in}(y_0) = \frac{1}{2(G_1 \pm G_{12})} \cdot \cos^2 \left( \frac{\pi}{L} y_0 \right), = Z_{in}(0) \cos^2 \cos^2 \left( \frac{\pi}{L} y_0 \right),$$  

(4.18)

where $Z_{in}(y_0)$ is the desired value of input impedance and $Z_{in}(0)$ is the impedance calculate in (4.17) [8].

Simulations of simple patch antenna have shown a BW of 2.2% on the chosen substrate. This value cannot achieve the desired value of 4.44%. Therefore BW improvements methods are studied in the following.

4.2 Bandwidth improvements

The first method improves the bandwidth by making $W = 2L$. With that it is possible to get a greater bandwidth and it is enough for the receiver antenna. However the radiation pattern changes and for the higher frequencies of the desired bandwidth the patch radiates to the sides instead of the broadside. Thus to other techniques are used to get the aimed bandwidth. The first technique uses two configuration called Non-radiating Edges Gap-Coupled Multiple Resonator broad-band Microstrip Antennas(NEGCOMA) and Radiating Edges Gap-Coupled Multiple Resonator broad-band Microstrip Antennas(REGCOMA) [9]. The second is the stacked antenna.
4.2.1 NEGCOMA and REGCOMA

The basic idea after NEGCOMA and REGCOMA is to place near to the original patch antenna (active patch) additional resonators. This creates a couple between the active patch and the additional resonators that allows to improve the bandwidth keeping the principal lobe to a broadside position. If the additional resonator is placed next to a radiating edge of the active patch it is called REGCOMA, meanwhile if it is placed next to non-radiating edge it is called NEGCOMA.

The size of the additional resonator is near the same of the original patch antenna. The width of the additional resonator is the same of the original patch. Thus the design parameters become the distance of the gap between the active patch and the additional resonator \(d_i\) and the length of the additional resonator \(L_i\).

For the gap-coupling between the two patch a loop in the Smith Chart appears. This loop depends on \(L_i\) and \(d_i\). When \(d_i\) increases the loop size becomes smaller and shifts to the right side of the Smith Chart. If \(d_i\) increases too much the loop disappears. On the other hand, when \(L_i\) decreases the loop shifts down and to the left of the Smith Chart.

When two additional resonators are placed the effects of changing \(d_i\) or \(l_i\) variate. Now there are two loops, one for each additional resonator. When one of the distances is reduced apart of increasing its loop size, the two loops become closer. This also happens when the difference between the lengths becomes smaller. Finally when the difference between the lengths is kept and the lengths are increased the loops shift upward and to the right of the Smith Chart.

With these considerations three additional resonators are placed, two of them in the non-radiating edges and one in the radiated edge, which is not feed, as shown figure 4.3. Then the structure is simulated in CST to optimize the different parameters. \(S_{11}\) is plotted at figure 4.5.

This structure has improved the bandwidth enough to be used as the receiver antenna. Also it keeps a broadside radiation pattern. However it has increased the width of the antenna at least three times, which makes the structure as big as sheet of paper DIN A4. Thus stacked antenna is proposed.

![Figure 4.3: NEGCOMA with one REGCOMA patch antenna.](image)

4.2.2 Stacked patch antennas

A stacked patch antenna is a structure that combines two dielectric substrates with patch antennas. One option is to place over the original patch antenna another substrate and then place a new patch antenna in
the top as shown figure 4.4. That makes at least a higher $h$ which means higher bandwidth. Also a coupling between the top and original patch appears. It is possible to change the coupling by tuning the distance $d$. In general the substrates could be different materials and have different thickness. To simplify in this thesis both substrates are taken equal to the filter’s one. The structure is simulated in CST. It is found that the bandwidth is large enough to be used as the receiver antenna. Also the radiation pattern is really similar to the simple patch antenna.

![Figure 4.4: Stack patch antenna.](image)

Figure 4.4: Stack patch antenna.

Figure 4.5 shows the $S_{11}$ for the different antenna configurations. How it shows, there are at least two suitable configurations: the NEGCOMA with one REGCOMA device and the stacked antenna. The first one has a bigger bandwidth than the stacked antenna. However it’s size is too big for the tag design. Thus, the receiver and the transmitter antennas are designed as stacked antenna.

![Figure 4.5: $S_{11}$ for different methods.](image)

Figure 4.5: $S_{11}$ for different methods.
4.3 Receiver antenna

Several simulations are made to adapt the receiver antenna as possible as can to the bandwidth required for the tag. Sizes of final design are represented in figure 4.7, also for the receiver antenna they have the following values: \( d = 1 \) mm, \( W_{\text{up}} = 74.4 \) mm, \( L_{\text{up}} = 62 \) mm, \( W_{\text{lower}} = 74.4 \) mm and \( L_{\text{lower}} = 62 \) mm. Also the lower patch has a inset feed patches have a notch to adapt the impedance with \( y_0 = 5 \) mm and a gap between the microstrip and patch sides of 4.83 mm. The upper patch has a notch placed in the same place where the lower patch has the inset feed. The size of the notch is \( W_{\text{notch}} = 14.49 \) mm and \( L_{\text{notch}} = 13 \) mm. The substrates thickness are the same \( t_{\text{up}} = t_{\text{lower}} = t = 1.575 \) mm.

Figure 4.8 compares the simulated and the measured \( S_{11} \). The bandwidth has shift to lower frequencies. However, the bandwidth is still in the specifications. Also it has larger bandwidth around 8.25 \% nearly four times the patch antenna's BW.

In order to measure the radiation pattern, the stacked antenna is placed in an anechoic chamber. With that measure it is possible to measure the E-plane and H-plane radiation patterns. They are represented with the stacked antenna's gain \( G_{\text{Stacked}} \) included in figures 4.9 and 4.10 respectively. The horn antenna ETS 3115 is used as reference to get the gain. This antenna has a gain of \( G_{\text{ref}} = 8.3 \) dB at 1.5 GHz. The gain of the stacked antenna could be determined using

\[
G_{\text{Stacked}} = P_{\text{Stacked}} - (P_{\text{ref}} - G_{\text{ref}}).
\]

(4.19)

where \( P_{\text{Stacked}} \) is the measured power of the stacked patch antenna and \( P_{\text{ref}} \) the measured power of the reference antenna. In this case the stacked antenna has a gain \( G_{\text{Stacked}} = 6.28 \) dB. The simulated one is 7.25 dB. So it is possible to say that the simulation results are quite accurate.

Also it is possible to plot the gain over the frequency to see the 8 \% BW over the frequency. Figure 4.11 shows it for different angles. As is expected at 0° it is maximum. Also it fits quiet well to the simulated results. However it shows that the measured bandwidth is greater than the simulated one as \( S_{11} \) shows. One possible

Figure 4.6: Fabricated antennas. Left transmitter antenna and right receiver antenna.
4.3 Receiver antenna

Figure 4.7: Dimensions of upper and lower patch.

Figure 4.8: Input matching of tag’s receiver antenna.

Figure 4.9: Receiver antenna: E-plane radiation pattern represented with gain.
explanation is that in the simulation the thickness of the glue that sticks the substrate is not contemplated. This thickness improves the general thickness and thus improves the bandwidth.

Figure 4.11: Receiver antenna gain vs frequency for different angles.

4.4 Transmitter antenna

To design the transmitter antenna is taken the same structure of the receiver antenna and the first approach is with the half of the size in all the parameters since the frequency range is twice. With the same representation of figure 4.7, its size values are for the distance \( d = 1 \text{ mm} \), \( W_{up} = 36.6 \text{ mm} \), \( L_{up} = 30.5 \text{ mm} \), \( W_{lower} = 36.6 \text{ mm} \) and \( L_{lower} = 30.5 \text{ mm} \). Also the lower patch has a inset feed patches have a notch to adapt the impedance with \( y_0 = 2 \text{ mm} \) and a gap between the microstrip and patch sides of 4.83 mm. The upper patch has a notch placed in the same place that the lower patch has the inset feed. The size of the notch is \( W_{notch} = 14.49 \text{ mm} \) and \( L_{notch} = 6 \text{ mm} \). The substrates thickness are the same \( t_{up} = t_{lower} = t = 1.575 \text{ mm} \).

In order to characterize the transmitter antenna, the same measurements of the receiver antenna are done. The measured \( S_{11} \) has again more bandwidth than the simulated as shows figure 4.12.

E-plane and H-plane radiation patterns are represented with the stacked antenna's gain \( G_{Stacked_Tx} \) included in figures 4.13 and 4.14 respectively. Also the same broadband horn antenna ETS 3115 is used as reference.
to get the gain. In this case the horn antenna has a gain of $G_{\text{ref}} = 9$ dB at 3.1 GHz. The gain of the stacked antenna could be determined using (4.19). In this case $G_{\text{Stacked}_{\text{tx}}} = 6.47$ dB while the simulated one is 7.37 dB. The measured BW is near 14.7%.

![Figure 4.12: $S_{11}$ from Transmitter antenna.](image)

4.4 Transmitter antenna
As in the receiver case the results are quite similar to the simulated ones. Also the bandwidth increase a bit. The ripples that appear in the gain appear also in the reference antenna measures. They could be a result of a weak in the source generator.

Figure 4.15: Transmitter antenna gain vs frequency for different angles.
5 Frequency doubler

A frequency doubler is an harmonic generator that produces at the output side a signal with two times the input frequency $f_{\text{out}} = 2 \cdot f_{\text{in}}$. In the ideal case at the output side would appear only the desired frequency as is shown in figure 5.1. However, usually undesired frequencies appear in the output side. These undesired frequencies are also harmonics of the input frequency which is called the fundamental frequency. Also this input frequency appears in the output side too. To generate this harmonics signals a non-linear device is needed as for example a diode, which its I-V is given for the Shockley diode equation

$$ I = I_s \left( e^{\frac{V_D}{nV_T}} - 1 \right), \quad (5.1) $$

where $I$ is the diode current, $I_s$ is the reverse bias saturation current, $V_D$ is the voltage across the diode, $V_T$ is the thermal voltage and finally $n$ is the ideality factor which determines the quality of the diode.

5.1 Zero-bias Schottky diode

A Schottky diode is a diode which has a low forward voltage drop and a fast switching action, since it is build as a metall-semiconductor junction. These properties allow it to work in environments which high frequency requirements and as a protector for solar panels or batteries. Zero-bias means that the diode works without a DC bias voltage, which allows to use it in situations without supply power as a passive chipless tag.

In this thesis is used the Avago Schottky diode HSMS-2850. In its datasheet appears the following equivalent circuit model and the encapsulate model as shown figure 5.2 and 5.3. Also it is indicated that the diode is design for input power under $-20 \text{ dBm}$. Also the parameters of the model are in the table 5.1. This model is implemented in Advance Design System of Agilent and matching networks are design to achieve the correct performance as a frequency doubler.
5.2 Matching Networks

In order to analyse the s-parameters of the diode, a simulation of the diode’s model is run. This simulations gets also the real and imaginary of the impedance of the diode for input and output frequencies. In this case the central input design frequency $f_{in} = 1.57 \text{ GHz}$

$$Z_{\text{diode input}}(f_{in}) = (109.8 - 561i) \Omega,$$  \hspace{1cm} (5.2)

and for the output side $f_{out} = 3.14 \text{ GHz}$

$$Z_{\text{diode output}}(f_{out}) = (83.7 - 281.3i) \Omega.$$  \hspace{1cm} (5.3)

Since the aim is to match the diode over the bandwidth of the system, the matching network should have the same in the input side and the double in the output. So the matching network should adapt different impedance at least for the input side from 1.54 GHz to 1.61 GHz and for the output side from 3.08 GHz to 3.22 GHz. Thus multiple $\frac{\lambda}{4}$ Chebyshev transformers are used. They can achieve large bandwidth for real impedances [10]. However their best performance is for real impedance they still can have a good performance for complex impedance. The first step is to add a series section to become the complex impedance a real one. Then the multiple $\frac{\lambda}{4}$ can be used.

Another handicap of the diode is the sensitivity of its impedance when the load is changed. Thus several iterations are done until a good matching networks are found. Also in the output side, an open stub is placed to filter the fundamental frequency and the second harmonic. Figure 5.4 shows a general equivalent circuit of these matching networks.
Figure 5.4: General schematic for the matching networks with $\frac{\lambda}{4}$ transformers.

First diode’s impedance of one port is placed in Smith Chart. Then a section is placed to get a real impedance. An finally the desired number of transformers are placed with its characteristics impedance. All this process is showed in figure 5.5.

Figure 5.5: Matching process step for the output side.

Simulation with this matching network is run and with the new values of the other port the process is repeated to create the second matching network. However, the first matching network needs to be recalculated again since the impedance seen from the first port has changed after placing the second matching network.

The final matching network is formed for two $\frac{\lambda}{4}$ in the input side and three in the output. Nevertheless the size of the matching is too big. In order to reduce the size an other substrate is used for the frequency doubler. The substrate is ROGERS RO 4003C with $\varepsilon_r = 3.55$ and thickness $t_2 = 0.813$ mm. Also some curves are made as shown figure 5.6 and 5.7. To check the performance an ADS Momentum simulation is run. This simulation shows that some length should be modified to achieve the desired performance. Figure 5.8 shows the fabricated prototype.
5.3 Measurements

The first measure are the S-parameters which are plotted in figure 5.9 and 5.10. They show that the measured frequency doubler has shifted to lower frequencies. So the matching now is worse than it was expected for the input frequency. In the output side the situation is a better since the output frequencies a less matching that the desire but just a few dB.

For the frequency doubler has more interest to measure its spectrum. That shows figure 5.11. In the spectrum the shift to lower frequencies appears too. Also the mismatching generates standing wave in the cables used to measure which are reflected in oscillations of the reading output power for the first harmonic. Despite of the shift in frequency and the mismatching, the fundamental in the output side is suppressed as it was desired. Also when the input power is lower the simulated results and the measured ones get closer. For example when the input power is $-40$ dBm the output power has nearly the same maximum value but in different frequencies.

Another interesting measure is to compare the matched diode with the unmatched diode. In this case, the matched diode is not really matched, but at least has better matching that the diode alone. Figure 5.12 shows
Figure 5.8: Fabricated frequency doubler.

Figure 5.9: Simulated and measured $S_{11}$ of the frequency doubler.

Figure 5.10: Simulated and measured $S_{21}$ of the frequency doubler.

Figure 5.11: Simulated and measured $P_{out}$ of the frequency doubler.

this comparison. This results are quite good, for the same input power the diode alone has near 20 dB less output power for the first harmonic and also between 20 dB and 40 dB more in the fundamental output.
An important parameter for a frequency doubler as for a mixer is the **Conversion loss** (CL). This parameter relates the output power for the first harmonic $P_{\text{out,1stHarm}}$ (dBm) and the input power of the fundamental $P_{\text{in,fund}}$ (dBm)

$$\text{CL(dB)} = P_{\text{in,fund}} - P_{\text{out,1stHarm}}.$$

(5.4)

The smaller this parameter the better is the frequency doubler.

Also is possible to define the term **Conversion gain** (CG) as

$$\text{CG(dB)} = P_{\text{out,1stHarm}} - P_{\text{in,fund}} = -\text{CL(dB)}.$$

(5.5)

Figure 5.14 shows the measured conversion loss of the frequency doubler over the input power of the fundamental for different frequencies. Again it is possible to see the shift in frequency since the minimum value of the conversion loss is for the maximum output power of the first harmonic. Meanwhile figure 5.13 shows the simulated and measured conversion gain over the frequency for different input powers. As is possible to see the maximum conversion gain was expected to be near 10 dB more in the simulation for $P_{\text{in}} = -20 \text{ dBm}$. Again when the input power becomes lower the measured value and the simulated become closer. One possible explanation for this fact is that the equivalent circuit model is better at lower input power since the diode is designed to be used under $-20 \text{ dBm}$. However both plots show that the lesser is $P_{\text{in}}$ the higher is the conversion loss or lower the conversion gain.
6 Measurement of the tag

After the design, fabrication and measurements of each part of the tag, the whole tag should be measured to prove if it is possible to codify the information in frequency and send it in the double frequency, which is the principal aim of this thesis.

In order to analyse the performance of the tag without antennas, wired measurements are done with the bandpass filter and the frequency doubler together. This wired measures could be used as a reference for the wireless ones, which are needed to prove the aim of this thesis.

6.1 Wired measurement

![Wired setup of temperature sensitive bandpass filter with frequency doubler.](image)

Figure 6.1: Wired setup of temperature sensitive bandpass filter with frequency doubler.

For the wired measurements the filter is connected at a signal generator which makes a sweep over the interest frequency band. The input side of the frequency doubler is connected to the band pass filter and the output side to spectrum analyser. From a pc it is possible to control, using GPIB port, both the spectrum analyser and the signal generator.

With this setup three different fixer cases of the DR are measured. The first one is using Rogers 3001 Bonding film, an other one is without any fixer and finally using cellophane tape. Figure 6.2 shows the results of this cases at room temperature of 22 °C. For the two last cases the measure is nearly the same the peak is at the same frequency as shows table 6.1. However they have different bandwidth and different transmitted power. This different power indicates that the losses with cellophane tape are greater than without any fixer. Also the $Q_L$ is greater for the case of no fixer. In other hand the shift of resonant frequency of 10 MHz could be explain for the fact of the Rogers film has a thickness of 0.038 mm and $\varepsilon_r = 2.28$ that is nearly the same $\varepsilon_r$ of the substrate. So for this case the total height is bigger so the resonate frequency should be lower.

The next measurement consists of temperature sweep for the cellophane fixer case. The results are shown in figure 6.3 and in the table 6.2. To increase the temperature an heating gun is used and an infrared camera is used to measure the DR's temperature. The frequency step of the signal generator is 1 MHz. The results show that the output power is nearly the same at this range of frequency. Also the frequency increment is nearly constant, near 11 MHz for an increment of 10 °C.
Figure 6.2: Wired measurement for different DR’s fixers for $P_{\text{in}} = -20 \, \text{dBm}$ for the setup of figure 6.1.

Table 6.1: Wired measurement for different DR’s fixers for $P_{\text{in}} = -20 \, \text{dBm}$ at $T = 22 \, ^\circ\text{C}$ for the setup of figure 6.1.

<table>
<thead>
<tr>
<th>Fixer</th>
<th>$P_{\text{out}}$(dBm)</th>
<th>$f_0$(GHz)</th>
<th>BW(MHz)</th>
<th>$Q_L$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cellophane</td>
<td>$-57.006$</td>
<td>3.122</td>
<td>4.3</td>
<td>727.30</td>
</tr>
<tr>
<td>No fixer</td>
<td>$-56.384$</td>
<td>3.122</td>
<td>3.4</td>
<td>909.40</td>
</tr>
<tr>
<td>Rogers 3001 Bonding Film</td>
<td>$-56.027$</td>
<td>3.112</td>
<td>3.9</td>
<td>796.98</td>
</tr>
</tbody>
</table>

Figure 6.3: Measure results for cellophane tape fixer for different temperatures for the setup of figure 6.1.
Table 6.2: Measure results for cellophane tape fixer for different temperatures for the setup of figure 6.1.

<table>
<thead>
<tr>
<th>T(°C)</th>
<th>$P_{\text{out}}$ (dBm)</th>
<th>$f_0$ (GHz)</th>
<th>BW (MHz)</th>
<th>$Q_L$</th>
</tr>
</thead>
<tbody>
<tr>
<td>22</td>
<td>−57.01</td>
<td>3.1220</td>
<td>4.3</td>
<td>727.30</td>
</tr>
<tr>
<td>32</td>
<td>−57.36</td>
<td>3.1310</td>
<td>3.3</td>
<td>954.11</td>
</tr>
<tr>
<td>38</td>
<td>−57.67</td>
<td>3.1370</td>
<td>3.4</td>
<td>930.79</td>
</tr>
<tr>
<td>42</td>
<td>−57.78</td>
<td>3.1410</td>
<td>4.3</td>
<td>723.38</td>
</tr>
<tr>
<td>48</td>
<td>−57.23</td>
<td>3.1470</td>
<td>4</td>
<td>777.23</td>
</tr>
</tbody>
</table>

6.2 Wireless measurement

Wireless measurements are necessary to prove the aim of this thesis. This wireless measurements are separated in unshielded measures and shielded ones. However the setup is similar at the one shown in figure 6.4. The tag is placed at 1 m from the antennas of the reading system. Also the heating gun is placed near the band pass filter to get it warmer.

Figure 6.4: Measurement setup.

6.2.1 Unshielded tag

For the unshielded tag there are three cases. The first one is shown in figure 6.5. This cases is referred in this thesis as Wireless Unshielded Setup 1 (WUS1). In this setup the whole parts of the tag are faced to the antennas of the reader. A temperature sweep measurements is realized for this case. The results are plotted in figure 6.6.
The results are not as good as expected. The noise floor should be under $-100 \text{ dBm}$ but the peak has less than 15 dB more power than a constant interference.

This interference signal comes from the filter and frequency doubler. Each one acts as antenna and has a contribution in this not desired signal.

To reduce this interference the second setup is used. In this case the filter and the frequency doubler are placed near the antennas of the reading system but aside of them to avoid their radiation beam. This situation is called in this thesis as Wireless Unshielded Setup 2 (WUS2). Again a temperature sweep measurement is done. The results are placed in figure 6.7 and table 6.4. In this case it is possible to see the noise floor. There is less power because the cables used to connect the tag antennas with the filter and the frequency doubler has near 4 dB losses. However the results are more as the expected ones. The difference between the peak power and the noise floor keeps more or less constant and over 20 dB. For WUS1 peak separation against the interference signal is not constant, it depends on the frequency.
The third case is a mix of the first two. The filter and the frequency doubler are turned 180° so their ground plane now is facing in front of the antennas of the reading system as in figure 6.8. The distance from the tag to the antennas in this case is 1.5 m. A temperature sweep measurement is made and the results are shown in figure 6.8 and table 6.5. This case is referred in this thesis as Wireless Unshielded Setup 3 (WUS3).

As in WUS2 case the results have improved respect to WUS1. Also in this measurement a greater temperature range was used and proves that the system works in the frequency range that is desired. The measurements

### Table 6.3: Results of WUS1.

<table>
<thead>
<tr>
<th>T(°C)</th>
<th>P_{out}(dBm)</th>
<th>f_0(GHz)</th>
<th>BW(MHz)</th>
<th>Q_L</th>
</tr>
</thead>
<tbody>
<tr>
<td>24</td>
<td>−69.95</td>
<td>3.1210</td>
<td>3.5</td>
<td>888.05</td>
</tr>
<tr>
<td>34</td>
<td>−72.10</td>
<td>3.1330</td>
<td>4.2</td>
<td>738.25</td>
</tr>
<tr>
<td>41</td>
<td>−72.14</td>
<td>3.1440</td>
<td>6.4</td>
<td>489.33</td>
</tr>
<tr>
<td>60</td>
<td>−72.55</td>
<td>3.1670</td>
<td>5.7</td>
<td>551.95</td>
</tr>
</tbody>
</table>

![Figure 6.7: Measured spectrum of WUS2.](image)

### Table 6.4: Results of WUS2.

<table>
<thead>
<tr>
<th>T(°C)</th>
<th>P_{out}(dBm)</th>
<th>f_0(GHz)</th>
<th>BW(MHz)</th>
<th>Q_L</th>
</tr>
</thead>
<tbody>
<tr>
<td>24</td>
<td>−86.94</td>
<td>3.1230</td>
<td>4.8</td>
<td>646.35</td>
</tr>
<tr>
<td>27</td>
<td>−87.68</td>
<td>3.1260</td>
<td>4</td>
<td>781.38</td>
</tr>
<tr>
<td>30</td>
<td>−88.21</td>
<td>3.1300</td>
<td>3.3</td>
<td>940.66</td>
</tr>
<tr>
<td>33</td>
<td>−89.57</td>
<td>3.1330</td>
<td>4.1</td>
<td>766.72</td>
</tr>
<tr>
<td>38</td>
<td>−89.34</td>
<td>3.1380</td>
<td>3.2</td>
<td>985.46</td>
</tr>
<tr>
<td>44</td>
<td>−89.68</td>
<td>3.1450</td>
<td>4</td>
<td>785.20</td>
</tr>
<tr>
<td>46</td>
<td>−88.79</td>
<td>3.1480</td>
<td>3.4</td>
<td>928.11</td>
</tr>
<tr>
<td>47</td>
<td>−88.57</td>
<td>3.1490</td>
<td>4.0</td>
<td>797.03</td>
</tr>
</tbody>
</table>
prove that it is possible to codify the temperature information in frequency domain with a passive chipless tag in combination with harmonic backscattering.

Figure 6.10 shows the dependence of the harmonic resonant frequency for all the setups. Equation (3.30) in section 3.2.3 defines a linear approximation of the dependence of $f_0$ with temperature. Thus a linear regression expression as

$$ f_0(T) = S_f T + f_0(0^\circ C) $$

(6.1)

is calculated with Matlab for each case in table 6.6. Also an expression function of $\Delta T$ is given. As they show, 1 $^\circ$C increase $f_0$ between 0.96 MHz and 1.29 MHz. So the system has a sensitivity near 1 MHz/$^\circ$C.

An other interesting factor is the reading power keeps inside a range of less than $-3$ dB when the temperature increases as shown figure 6.11. However the loaded Q-Factor $Q_L$ of the global system changes strongly

6.2 Wireless measurement
Table 6.5: Results of WUS3.

<table>
<thead>
<tr>
<th>T(°C)</th>
<th>P_{out}(dBm)</th>
<th>f_{0}(GHz)</th>
<th>BW(MHz)</th>
<th>Q_L</th>
</tr>
</thead>
<tbody>
<tr>
<td>22.00</td>
<td>−80.25</td>
<td>3.1205</td>
<td>3.3</td>
<td>952.26</td>
</tr>
<tr>
<td>37.00</td>
<td>−80.57</td>
<td>3.1415</td>
<td>2.6</td>
<td>1205.73</td>
</tr>
<tr>
<td>45.00</td>
<td>−81.48</td>
<td>3.1560</td>
<td>3.5</td>
<td>907.40</td>
</tr>
<tr>
<td>51.00</td>
<td>−81.43</td>
<td>3.1600</td>
<td>3.6</td>
<td>875.59</td>
</tr>
<tr>
<td>67.00</td>
<td>−81.11</td>
<td>3.1815</td>
<td>5.5</td>
<td>575.56</td>
</tr>
<tr>
<td>72.00</td>
<td>−81.38</td>
<td>3.1840</td>
<td>5.6</td>
<td>566.95</td>
</tr>
<tr>
<td>80.00</td>
<td>−78.55</td>
<td>3.1975</td>
<td>4.8</td>
<td>669.88</td>
</tr>
<tr>
<td>99.00</td>
<td>−78.60</td>
<td>3.2195</td>
<td>4.6</td>
<td>692.42</td>
</tr>
</tbody>
</table>

Figure 6.10: Measured frequency over the temperature for the different unshielded setup.

with this increase. A possible explanation is that this behaviour is caused for two different facts. Since the frequency doubler has shift the central frequency and its matching has shift also, it is possible that the power which comes from the filter has different reflection coefficient over the frequency so sometimes there are more frequencies which are matched to the diode and sometimes is less. Also the frequency doubler has greater conversion loss for the lower power so it could filter the peak making it sharper. Then the bandwidth decrease and the unloaded Q-Factor Q_L increase. This also could explain why the Q_L of the whole tag is larger than that of the filter only.

6.2.2 Shielded Tag

The shielded tag is the tag with the filter and the frequency doubler shielded with some metal cover as they show figure 6.13 and 6.14. In this case the shield is made of aluminium foil.
Table 6.6: Lineal regression for unshielded measures.

<table>
<thead>
<tr>
<th>Measured case</th>
<th>Lineal regression</th>
<th>$f(\Delta T)$ centred at $T = 22^\circ C$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wired</td>
<td>$f_0(T) = 9.6556 \text{ MHz/}^\circ C \cdot T + 3.1005 \text{ GHz}$</td>
<td>$f(\Delta T) = 3.1217 \text{ GHz} \cdot (1 + 308.67 \text{ ppm/}^\circ C \cdot \Delta T)$</td>
</tr>
<tr>
<td>WUS1</td>
<td>$f_0(T) = 12.865 \text{ MHz/}^\circ C \cdot T + 3.0901 \text{ GHz}$</td>
<td>$f(\Delta T) = 3.1184 \text{ GHz} \cdot (1 + 412.55 \text{ ppm/}^\circ C \cdot \Delta T)$</td>
</tr>
<tr>
<td>WUS2</td>
<td>$f_0(T) = 11.264 \text{ MHz/}^\circ C \cdot T + 3.0958 \text{ GHz}$</td>
<td>$f(\Delta T) = 3.1206 \text{ GHz} \cdot (1 + 360.96 \text{ ppm/}^\circ C \cdot \Delta T)$</td>
</tr>
<tr>
<td>WUS3</td>
<td>$f_0(T) = 12.688 \text{ MHz/}^\circ C \cdot T + 3.095 \text{ GHz}$</td>
<td>$f(\Delta T) = 3.1229 \text{ GHz} \cdot (1 + 406.29 \text{ ppm/}^\circ C \cdot \Delta T)$</td>
</tr>
</tbody>
</table>

Figure 6.11: Measured normalized output power over the temperature for the different unshielded set up.

Figure 6.12: Measured $Q_L$ over the temperature for the different unshielded set up.

Figure 6.13: Shielded Tag.

Figure 6.14: Shielded filter and shielded frequency doubler.

With this configuration two different measures are made. First a sweep temperature measurement is made and represented in figure 6.15 and table 6.7. However, the aluminium foil does not allow to measure the DR temperature. That is the reason for it appears as increments $\Delta T_i$. Furthermore if the sensitivity of the unshielded DR is taken the temperatures could be calculated by subtracting to the frequency of the $\Delta T_i$ the reference frequency all in MHz. The difference would be $\Delta T_i$. On other hand, the values of $Q_L$ are higher than in the unshielded one as it is in the simulated results of the filter.

After the temperature measurement a distance one is made. The tag is moved from 1 m to 2.8 m. The results are in figure 6.16 and table 6.8. What they show is that $f_0$ is not constant with the distance. However his shift is less than the equivalent shift for 1°C. Figure 6.17 shows the measured harmonic resonant frequency over the distance. Also it show the mean value of $f_0$ and the mean ± standard deviation. Also figure 6.18
Figure 6.15: Measured spectrum of shielded tag for different temperatures.

Table 6.7: Shielded tag’s results for different temperatures.

<table>
<thead>
<tr>
<th>T(°C)</th>
<th>$P_{out}$(dBm)</th>
<th>$f_0$(GHz)</th>
<th>BW(MHz)</th>
<th>$Q_L$</th>
</tr>
</thead>
<tbody>
<tr>
<td>22</td>
<td>−70.14</td>
<td>3.1245</td>
<td>2</td>
<td>1562.12</td>
</tr>
<tr>
<td>(22+ΔT_1)</td>
<td>−67.28</td>
<td>3.1265</td>
<td>2.5</td>
<td>1231.63</td>
</tr>
<tr>
<td>(22+ΔT_2)</td>
<td>−66.88</td>
<td>3.1315</td>
<td>2.4</td>
<td>1302.69</td>
</tr>
<tr>
<td>(22+ΔT_3)</td>
<td>−64.39</td>
<td>3.1370</td>
<td>2.7</td>
<td>1151.34</td>
</tr>
<tr>
<td>(22+ΔT_4)</td>
<td>−64.42</td>
<td>3.1425</td>
<td>3</td>
<td>1047.41</td>
</tr>
<tr>
<td>(22+ΔT_5)</td>
<td>−64.56</td>
<td>3.1465</td>
<td>3</td>
<td>1033.85</td>
</tr>
<tr>
<td>(22+ΔT_6)</td>
<td>−64.84</td>
<td>3.1490</td>
<td>3.5</td>
<td>909.94</td>
</tr>
<tr>
<td>(22+ΔT_7)</td>
<td>−64.70</td>
<td>3.1520</td>
<td>3.4</td>
<td>918.11</td>
</tr>
</tbody>
</table>

shows the error of each measure if the mean is taken as the real value of $f_0$. In this case the error is less than 0.35 MHz. It is possible to define the parameter absolute error as

$$e_{\text{Frequency}} = |f_{\text{mean}} - f_{\text{measured}}|,$$  \hspace{1cm} (6.2)

with is plotted in figure 6.18. It shows that the error is less than 0.35 MHz.

Finally a plot of the received power over the distance is shown in figure 6.19. The results shows that for a transmitted power from the reading system of 10 dBm is possible to read 2.8 m with more than −92 dBm.

6.2 Wireless measurement
Figure 6.16: Measured spectrum of shielded tag for different distances $d$.

Table 6.8: Results of shielded tag for different distances $d$.

<table>
<thead>
<tr>
<th>$d$(m)</th>
<th>$P_{out}$(dBm)</th>
<th>$f_0$(GHz)</th>
<th>BW(MHz)</th>
<th>$Q_L$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.00</td>
<td>−69.67</td>
<td>3.1254</td>
<td>1.9</td>
<td>1658.49</td>
</tr>
<tr>
<td>1.20</td>
<td>−72.00</td>
<td>3.1252</td>
<td>1.8</td>
<td>1778.02</td>
</tr>
<tr>
<td>1.40</td>
<td>−77.39</td>
<td>3.1252</td>
<td>1.7</td>
<td>1804.57</td>
</tr>
<tr>
<td>1.60</td>
<td>−78.53</td>
<td>3.1252</td>
<td>1.6</td>
<td>1922.32</td>
</tr>
<tr>
<td>1.80</td>
<td>−79.78</td>
<td>3.1252</td>
<td>1.5</td>
<td>2152.88</td>
</tr>
<tr>
<td>2.00</td>
<td>−84.14</td>
<td>3.1250</td>
<td>1.5</td>
<td>2145.12</td>
</tr>
<tr>
<td>2.20</td>
<td>−83.64</td>
<td>3.1248</td>
<td>1.7</td>
<td>1862.56</td>
</tr>
<tr>
<td>2.40</td>
<td>−83.67</td>
<td>3.1248</td>
<td>1.5</td>
<td>2135.23</td>
</tr>
<tr>
<td>2.60</td>
<td>−86.07</td>
<td>3.1248</td>
<td>1.6</td>
<td>2006.99</td>
</tr>
<tr>
<td>2.80</td>
<td>−91.63</td>
<td>3.1250</td>
<td>1.4</td>
<td>2190.34</td>
</tr>
</tbody>
</table>

Figure 6.17: Measured harmonic resonant frequency over distance of shielded tag.
Figure 6.18: Absolute error of measured frequency over the distance.

Figure 6.19: Received power over distance of shielded tag.
7 Conclusion

A harmonic radar temperature sensing technique for a chipless tag has been presented in this thesis. The wireless measurements have successfully proved the concept and a successful determination of the harmonic resonant frequency has been achieved. The tag codifies the temperature sensing in the resonant frequency of a dielectric resonator (DR) and then the frequency doubler translate it to the double resonant frequency. Then the tag sends it to the reader system. Both the tag and the reader system are provided with two antennas, one at fundamental band from 1.5 GHz to 1.64 GHz and the other at the first harmonic band from 3 GHz to 3.28 GHz.

This harmonic temperature sensor can sense temperatures below 20 °C and near 100 °C. Also it has measured up to a distance of 2.8 m with a transmit power of the reader system of 10 dBm. The tag shows an absolute sensitivity 1.3 MHz/K. It consists of two stacked patch antennas, a temperature-dependent DR bandpass filter and a harmonic generator. The stacked antennas show a gain over 6 dB with a 3 dB bandwidth of 8%. The temperature sensitive bandpass filter has achieved at least a loaded Q-Factor of 470 and an insertion loss of 4.5 dB. The harmonic generator has been optimized with two matching networks (each one for the input and output side) to achieve a conversion losses near 20 dB. However, it is found that the frequency doubler also called harmonic generator has shifted to a lower frequencies from the desired one and the conversion losses has increased, but it is still used since it has near 20 dB more in the first harmonic than the Schottky diode alone. It reduces the fundamental frequency in the output signal too.

Also has been found that the frequency doubler and the filter act as antennas and introduce undesired signals that effect the performance of the tag. To solve this problem two suitable configurations have been presented. The first one consists in shielding the filter and the frequency doubler. That allows to have greater loaded Q-Factor $Q_L$ with the same output power. Nevertheless the DR is closed in a box. That can create a high different of temperature inside the box and outside. A possible way to improve this shielded sensor could be to make some holes in the metal box around the DR that allow the pass of the air. However to avoid undesired radiation this holes should be small. The second method consists of facing the ground plane of the filter and the mixer directly to the reader system's antennas.

Furthermore the lower fundamental frequency makes the tag too large. A way to make the tag smaller could be use the DR as antenna depending on temperature, which means that the antenna’s frequency shift with the temperature. Other possibility is to reduce the size of the DR keeping its aspect ratio. This will increase the resonant frequency of the DR. Also it is possible to use a substrate with higher $\varepsilon_r$, which will reduce the whole size of the system. An other improve could be the use of different harmonics generators that allows the system to have more than one sensor in the same room. Also if ceramic substrate is used it is possible to achieve higher temperature range.
A Layouts
Bibliography


List of Figures

2.1 a) Traditional backscattering sensing. b) Harmonic sensing ............................................. 5
2.2 Schematic harmonic radar temperature sensor. ................................................................. 6

3.1 DR placed on substrate. ........................................................................................................ 8
3.2 DR between metallic planes. ................................................................................................ 10
3.3 $Q_{\text{rad}}$ in function of $\frac{h_{\text{dr}}}{d}$ with $r_{\text{ref}}$ normalized for isolated DR with mode $\text{TE}_{011+\delta}$ ................................................................. 12
3.4 Different cases of $\Delta \bar{f}$ with the same $\Delta T_0$. ............................................................. 12
3.5 Filter configurations. (a) Type I. (b) Type II. ................................................................. 13
3.6 S-parameter for DR placed on substrate where $d_1 = d_2 = d$ and $t = 0.8$ mm. In the left, s-parameters from ideal case. In the middle, s-parameters with only DR losses. In the right, s-parameters with all losses. ................................................................. 16
3.7 Power for DR placed on substrate where $d_1 = d_2 = d$ and $t = 0.8$ mm. In the left, $P_{\text{loss}}$ for the different losses cases. In the middle, $P_{\text{trans}}$ for the different losses cases. In the right, $P_{\text{ref}}$ for the different losses cases. ................................................................. 17
3.8 S-parameter for shielded DR placed on substrate where $d_1 = d_2 = d$ and $t = 0.8$ mm. In the left, s-parameters from ideal case. In the middle, s-parameters with only DR losses. In the right, s-parameters with all losses. ................................................................. 18
3.9 Power for shielded DR placed on substrate where $d_1 = d_2 = d$ and $t = 0.8$ mm. In the left, $P_{\text{loss}}$ for the different losses cases. In the middle, $P_{\text{trans}}$ for the different losses cases. In the right, $P_{\text{ref}}$ for the different losses cases. ................................................................. 18
3.10 Loaded Quality factor $Q_L$ over $d$ for the shielded and unshielded cases. ...................... 19
3.11 S-parameters for $t = 1.575$ mm, unshielded case. ........................................................... 19
3.12 Normalized Power over $d$ for $t = 1.575$ mm, unshielded case. ........................................ 20
3.13 Loaded Q-Factor $Q_L$ for $t = 1.575$ mm, unshielded cas. .................................................. 20
3.14 $S_{21}$ for different sizes of glue and tin fix. ........................................................................... 21
3.15 Simulated s-parameters for mesh 4 and 6. ........................................................................... 21
3.16 $S_{21}$ for the cellophane and no fixer at $T = 22^\circ$C. .............................................................. 22
3.17 Shift of $f_0$ for different fixers. .......................................................................................... 22
3.18 $f_0$ over $T$ ............................................................................................................................ 23

4.1 Rectangular patch antenna. ............................................................... 25
4.2 Radiation pattern of patch antenna simulated with CST. .................................................. 25
4.3 NEGCOMA with one REGCOMA patch antenna. ............................................................ 27
4.4 Stack patch antenna. .......................................................................................................... 28
4.5 $S_{11}$ for different methods. ............................................................................................... 28
4.6 Fabricated antennas. Left transmitter antenna and right receiver antenna. ...................... 29
4.7 Dimensions of upper and lower patch. .............................................................................. 30
4.8 Input matching of tag’s receiver antenna. .......................................................................... 30
4.9 Receiver antenna: E-plane radiation pattern represented with gain. .................................. 30
4.10 Receiver antenna: H-plane radiation pattern represented with gain. .................................. 31
4.11 Receiver antenna gain vs frequency for different angles. .................................................. 31
4.12 $S_{11}$ from Transmitter antenna. ......................................................................................... 32
4.13 Transmitter antenna: E-plane radiation pattern represented with gain. ........................... 32
4.14 Transmitter antenna: H-plane radiation pattern represented with gain. ........................... 32
4.15 Transmitter antenna gain vs frequency for different angles. ........................................ 33

5.1 Desired performance of the frequency doubler. .......................................................... 34
5.2 Schottky diode’s equivalent circuit model. ................................................................. 35
5.3 Schottky diode’s encapsulate circuit model. ............................................................... 35
5.4 General schematic for the matching networks with $\frac{1}{4}$ transformers. ....................... 36
5.5 Matching process step for the output side. ................................................................. 36
5.6 ADS schematic of the matching network ................................................................. 37
5.7 Layout of the matching network .............................................................................. 37
5.8 Fabricated frequency doubler. ................................................................................... 38
5.9 Simulated and measured $S_{11}$ of the frequency doubler ............................................ 38
5.10 Simulated and measured $S_{21}$ of the frequency doubler ............................................ 38
5.11 Simulated and measured $P_{out}$ of the frequency doubler .......................................... 38
5.12 $P_{out}$ for the fundamental and the first harmonic for matched and unmatched diode. ....... 39
5.13 Simulated and measured conversion gain over frequency for different $P_{in}$. ............... 39
5.14 Measured conversion loss over $P_{in}$ for different frequencies. .................................. 39

6.1 Wired setup of temperature sensitive bandpass filter with frequency doubler. ............... 40
6.2 Wired measurement for different DR's fixers for $P_{in} = -20$ dBm for the setup of figure 6.1 . 41
6.3 Measure results for cellophane tape fixer for different temperatures for the setup of figure 6.1. 41
6.4 Measurement setup. ......................................................................................... 42
6.5 Wireless Unshielded Setup 1 (WUS1). ....................................................................... 43
6.6 Measured spectrum of WUS1. .................................................................................. 43
6.7 Measured spectrum of WUS2. .................................................................................. 44
6.8 WUS3 measurement setup. ..................................................................................... 45
6.9 Measured spectrum of WUS3 for different temperatures. .......................................... 45
6.10 Measured frequency over the temperature for the different unshielded setup. ............. 46
6.11 Measured normalized output power over the temperature for the different unshielded set up. 47
6.12 Measured $Q_L$ over the temperature for the different unshielded set up. ....................... 47
6.13 Shielded Tag. ........................................................................................................ 47
6.14 Shielded filter and shielded frequency doubler. ......................................................... 47
6.15 Measured spectrum of shielded tag for different temperatures. ................................. 48
6.16 Measured spectrum of shielded tag for different distances $d$ ..................................... 49
6.17 Measured harmonic resonant frequency over distance of shielded tag. ...................... 49
6.18 Absolute error of measured frequency over the distance. ........................................ 50
6.19 Received power over distance of shielded tag. ....................................................... 50
List of Tables

3.1 Equivalent modes from isolated DR to DR on metal plane. .............................. 11
3.2 Theoretical value of resonant frequency from the sample for the different methods. ....... 15
3.3 Measured value of resonant frequency for different values of T for the film fixer. ......... 23

5.1 EqParameters of the equivalent circuit model of HSMS-2850 Schottky diode. ............ 35

6.1 Wired measurement for different DR's fixers for $P_{in} = -20$ dBm at $T = 22$ °C for the setup of figure 6.1. ................................................................. 41
6.2 Measure results for cellophane tape fixer for different temperatures for the setup of figure 6.1. 42
6.3 Results of WUS1. .............................. 44
6.4 Results of WUS2. .............................. 44
6.5 Results of WUS3. .............................. 46
6.6 Lineal regression for unshielded measures. .................................................. 47
6.7 Shielded tag's results for different temperatures. ........................................ 48
6.8 Results of shielded tag for different distances $d$. .............................. 49