Design and Performance Assessment of Ground Plane Boosters with a Bar Form Factor

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Abstract

The progress that technology has experienced in the last few decades has largely improved electronic devices, particularly mobile phones, which have clearly evolved from having basic features to almost PC-like software and high speed connectivity. Every year, hundreds of thousands of cell phones are sold around the world. Customers demand both feature phones and smartphones, names meaning how low or high the processing and multimedia capabilities of the phone are, which ultimately sets the price tag. Yet, regardless of phones being one type or the other, they all require connectivity to cellular services. Providing wireless communication traditionally required antennas whose size depend on the wavelength of the frequency in which the cellular standard was deployed; usually said sizes are quite large with respect to the dimensions of a phone. However, antennas evolved from external to internal, and from singleband to multiband. This made the geometry of antennas change from very simple to very complex shapes, making it possible to communicate in several bands with an antenna being completely inside the handset. The ground plane booster antenna technology is able to provide connectivity without requiring an antenna by advantageously using the ground plane of the device, and still feature a radiating system comprised of boosters which is completely internal, multiband, and, moreover, does not require custom complex geometry designs. Ground plane boosters are small-sized elements that enable the ground plane as the radiator of a device and, hence, act like an antenna. In line with the trend of electronic devices becoming thinner, and in spite of boosters already being very small, a new form factor for ground plane boosters is studied and developed in this thesis such that their RF performance is satisfactory according to the standards in the cellular industry, specifically in 2G, 3G and 4G services. For the design of a new booster form factor, the effect each booster dimension has on the radioelectric performance of the radiating structure is first studied. Then, shape and dimensions of several booster candidates are set and their RF performance is simulated. The form factor that achieves the best results according to the simulations is prototyped together with different matching networks, and reflection coefficient and antenna efficiency are both measured. In last place, a final triple-port printed circuit board with boosters with the new form factor is presented along with the matching networks synthesized for operation in the main 2G, 3G and 4G frequency bands.
# Contents

Abstract

1. Introduction
   1.1. Objectives and Scope ........................................... 2
   1.2. Structure of the Thesis ....................................... 3
   1.3. Work Methodology ............................................. 4
      1.3.1. Documentation ........................................... 4
      1.3.2. Simulations ............................................. 5
      1.3.3. Prototyping and Measuring ............................... 6
   1.4. Patents ....................................................... 7

2. Basic Concepts Related to Antennas .................................. 9
   2.1. Reflection Coefficient ...................................... 9
   2.2. Antenna Bandwidth .......................................... 10
   2.3. Quality Factor Q of Antennas and Inherent Bandwidth .... 10
   2.4. Antenna and Radiation Efficiencies ......................... 12
   2.5. Quality Factor Q of Lumped Components ..................... 13

3. Ground Plane Boosters .................................................. 15
   3.1. Introduction .................................................. 15
   3.2. Theory ........................................................ 16
   3.3. Performance .................................................. 19

4. Matching Networks for Ground Plane Boosters ......................... 23
   4.1. Introduction .................................................. 23
   4.2. Mismatch Loss Based Designs ................................. 24
   4.3. Transducer Power Gain Based Designs ....................... 25
   4.4. Matching Networks for Ground Plane Boosters ............... 28
   4.5. Conclusions .................................................. 32

5. Ground Plane Boosters with a Bar Form Factor ......................... 33
   5.1. Introduction .................................................. 33
   5.2. Study of Form Factors ...................................... 34
      5.2.1. Starting Point for Design and Reference Values .... 34
      5.2.2. Changes in Length and Width Dimensions of Booster .. 37
List of Figures

1.1. Photo of Fractus® headquarters ........................................ 3
1.2. Mentor Graphics® IE3D™ .................................................. 5
1.3. Microwave Office® ........................................................... 6
1.4. Photo of Fractus® chemical laboratory .................................. 6
1.5. Photos of network analyzer and anechoic chamber .................... 7
2.1. Q factors of commercial capacitor and inductor versus frequency .... 14
2.2. ESR of commercial capacitor and inductor versus frequency ......... 14
3.1. Eigenvalue and modal significance representations of ground plane modes . 17
3.2. Ground plane boosters exciting a radiation mode in two frequency regions . 18
3.3. Simulated impedance of an exemplary unmatched radiating structure . . 20
3.4. Matching networks for an exemplary matched radiating structure ....... 20
3.5. Simulated impedances of an exemplary matched radiating structure .... 21
3.6. Simulated S_{11} parameter of an exemplary matched radiating structure .... 21
3.7. Measured S_{11} parameter of an exemplary matched radiating structure .... 22
3.8. Measured \eta_a and computed \eta_r of an exemplary matched radiating structure .... 22
4.1. Matching network for transducer power gain exemplification ............ 26
4.2. Simulated input impedance using exemplary matching networks ........ 27
4.3. Simulated S_{21} parameter using exemplary matching networks .......... 27
4.4. Schematic of ladder matching network for LTE700 ..................... 29
4.5. Schematic of broadband matching network for LFR .................... 29
4.6. Simulated S_{21} parameter considering first exemplary lossy BB-MN ....... 31
4.7. Simulated S_{21} parameter considering second exemplary lossy BB-MN ....... 31
4.8. Schematic of L-topology matching network for HFR ................... 32
4.9. Schematic of T-topology matching network for HFR ................... 32
5.1. Definition of booster dimensions ........................................ 35
5.2. Simulated inherent bandwidth versus booster’s length dimension ........ 38
5.3. Simulated inherent bandwidth versus booster’s width dimension .......... 39
5.4. Simulated inherent bandwidth versus booster’s gap dimension ........... 40
5.5. Simulated inherent bandwidth versus booster’s gap and width dimensions .... 41
5.6. Simulated input impedance of a 5x5x5mm³ booster in LFR ............. 44
5.7. Simulated input impedance of a 3.2x3.2x7mm³ booster in LFR ............ 45
5.8. Simulated input impedance of a 3.2x3.2x10mm$^3$ booster in LFR . . . . . . . 46
5.9. Comparison of simulated $S_{11}$ parameter for three boosters in LFR . . . . . . 46
5.10. Simulated input impedance of a 5x5x5mm$^3$ booster in HFR . . . . . . . 47
5.11. Simulated input impedance of a 3.2x3.2x7mm$^3$ booster in HFR . . . . . . 48
5.12. Simulated input impedance of a 3.2x3.2x10mm$^3$ booster in HFR . . . . . . 49
5.13. Comparison of simulated $S_{11}$ parameter for three boosters in HFR . . . . . 49
5.14. Simulated input impedance of a 5x5x5mm$^3$ booster in LTE700 . . . . . . 50
5.15. Simulated input impedance of a 3.2x3.2x7mm$^3$ booster in LTE700 . . . . . 51
5.16. Simulated input impedance of a 3.2x3.2x10mm$^3$ booster in LTE700 . . . . 51
5.17. Comparison of simulated $S_{11}$ parameter for three boosters in LTE700 . . . 52
5.18. Photo of brass-made boosters sized 3.2x3.2x10mm$^3$ . . . . . . . . . . . . . 54
5.19. AutoDesk® AutoCAD®'s 3D perspective of the triple-port PCB's layout . . 55
5.20. Photo of a single-port PCB with 3 pads for boosters sized 3.2x3.2x10mm$^3$ . 55
5.21. Measured input impedance of a 3.2x3.2x10mm$^3$ booster . . . . . . . . . . . 56
5.22. Measured input impedance of booster with first BB-MN in LFR (1-port) . . . 57
5.23. Measured $S_{11}$ and $\eta_a$ of booster with first BB-MN in LFR (1-port) . . . 57
5.24. Measured input impedance of booster with second BB-MN in LFR (1-port) . 58
5.25. Measured $S_{11}$ and $\eta_a$ of booster with second BB-MN in LFR (1-port) . . 58
5.26. Measured input impedance of booster with third BB-MN in LFR (1-port) . . 60
5.27. Measured $S_{11}$ and $\eta_a$ of booster with third BB-MN in LFR (1-port) . . . 60
5.28. Measured input impedance of booster with fourth BB-MN in LFR (1-port) . 61
5.29. Measured $S_{11}$ and $\eta_a$ of booster with fourth BB-MN in LFR (1-port) . . 61
5.30. Measured input impedance of booster with fifth BB-MN in LFR (1-port) . . . 62
5.31. Measured $S_{11}$ and $\eta_a$ of booster with fifth BB-MN in LFR (1-port) . . . 62
5.32. Measured input impedance of booster with T-MN in HFR (1-port) . . . . . . 64
5.33. Measured $S_{11}$ and $\eta_a$ of booster with T-MN in HFR (1-port) . . . . . . 65
5.34. Measured input impedance of booster with first L-MN in HFR (1-port) . . . 65
5.35. Measured $S_{11}$ and $\eta_a$ of booster with first L-MN in HFR (1-port) . . . 66
5.36. Measured input impedance of booster with second L-MN in HFR (1-port) . 66
5.37. Measured $S_{11}$ and $\eta_a$ of booster with second L-MN in HFR (1-port) . . 67
5.38. Measured input impedance of booster with third L-MN in HFR (1-port) . . 67
5.39. Measured $S_{11}$ and $\eta_a$ of booster with third L-MN in HFR (1-port) . . . 68
5.40. Measured input impedance of a 3.2x3.2x20mm$^3$ booster . . . . . . . . . . . 69
5.41. Measured input impedance of booster with first ladder MN in LTE700 (1-port) . 70
5.42. Measured $S_{11}$ and $\eta_a$ of booster with first ladder MN in LTE700 (1-port) . 70
5.43. Measured input impedance of booster with sec. ladder MN in LTE700 (1-port) . 71
5.44. Measured $S_{11}$ and $\eta_a$ of booster with second ladder MN in LTE700 (1-port) . 71
5.45. Photo of the final triple-port matching networks . . . . . . . . . . . . . . . . 72
5.46. Photo of the final triple-port PCB including four boosters sized 3.2x3.2x10mm$^3$ . 73
5.47. Measured input impedance of final bar-shaped booster in LFR (3-port) . . . 73
5.48. Measured $S_{11}$ and $\eta_a$ of final bar-shaped booster in LFR (3-port) . . . . 74
5.49. Measured input impedance of non-final bar-shaped booster in HFR (3-port) . 75
5.50. Measured $S_{11}$ and $\eta_a$ of non-final bar-shaped booster in HFR (3-port) . . 75
5.51. Measured input impedance of final bar-shaped booster in HFR (3-port) . . . 76
<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.52</td>
<td>Measured $S_{11}$ and $\eta_a$ of final bar-shaped booster in HFR (3-port)</td>
<td>77</td>
</tr>
<tr>
<td>5.53</td>
<td>Measured input impedance of final bar-shaped booster in LTE700 (3-port)</td>
<td>77</td>
</tr>
<tr>
<td>5.54</td>
<td>Measured $S_{11}$ and $\eta_a$ of final bar-shaped booster in LTE700 (3-port)</td>
<td>78</td>
</tr>
<tr>
<td>5.55</td>
<td>Measured $S_{12}$ of final bar-shaped boosters in triple-port configuration</td>
<td>78</td>
</tr>
<tr>
<td>5.56</td>
<td>Main measured 2D radiation pattern cuts</td>
<td>79</td>
</tr>
<tr>
<td>6.1</td>
<td>Photos of Fractus® bar-shaped and cube-shaped booster products</td>
<td>82</td>
</tr>
<tr>
<td>A.1</td>
<td>Matching network for 5x5x5mm$^3$ booster simulation in LFR</td>
<td>91</td>
</tr>
<tr>
<td>A.2</td>
<td>Matching network for 3.2x3.2x7mm$^3$ booster simulation in LFR</td>
<td>91</td>
</tr>
<tr>
<td>A.3</td>
<td>Matching network for 3.2x3.2x10mm$^3$ booster simulation in LFR</td>
<td>92</td>
</tr>
<tr>
<td>A.4</td>
<td>Matching network for 5x5x5mm$^3$ booster simulation in HFR</td>
<td>92</td>
</tr>
<tr>
<td>A.5</td>
<td>Matching network for 3.2x3.2x7mm$^3$ booster simulation in HFR</td>
<td>92</td>
</tr>
<tr>
<td>A.6</td>
<td>Matching network for 3.2x3.2x10mm$^3$ booster simulation in HFR</td>
<td>93</td>
</tr>
<tr>
<td>A.7</td>
<td>Matching network for 5x5x10mm$^3$ booster simulation in LTE700</td>
<td>93</td>
</tr>
<tr>
<td>A.8</td>
<td>Matching network for 3.2x3.2x14mm$^3$ booster simulation in LTE700</td>
<td>93</td>
</tr>
<tr>
<td>A.9</td>
<td>Matching network for 3.2x3.2x20mm$^3$ booster simulation in LTE700</td>
<td>94</td>
</tr>
<tr>
<td>B.1</td>
<td>First BB-MN for 3.2x3.2x10mm$^3$ booster in LFR, 1-port config</td>
<td>95</td>
</tr>
<tr>
<td>B.2</td>
<td>Second BB-MN for 3.2x3.2x10mm$^3$ booster in LFR, 1-port config</td>
<td>96</td>
</tr>
<tr>
<td>B.3</td>
<td>Third BB-MN for 3.2x3.2x10mm$^3$ booster in LFR, 1-port config</td>
<td>96</td>
</tr>
<tr>
<td>B.4</td>
<td>Fourth BB-MN for 3.2x3.2x10mm$^3$ booster in LFR, 1-port config</td>
<td>96</td>
</tr>
<tr>
<td>B.5</td>
<td>Fifth BB-MN for 3.2x3.2x10mm$^3$ booster in LFR, 1-port config</td>
<td>97</td>
</tr>
<tr>
<td>B.6</td>
<td>T-MN for 3.2x3.2x10mm$^3$ booster in HFR, 1-port config</td>
<td>97</td>
</tr>
<tr>
<td>B.7</td>
<td>First L-MN for 3.2x3.2x10mm$^3$ booster in HFR, 1-port config</td>
<td>97</td>
</tr>
<tr>
<td>B.8</td>
<td>Second L-MN for 3.2x3.2x10mm$^3$ booster in HFR, 1-port config</td>
<td>98</td>
</tr>
<tr>
<td>B.9</td>
<td>Third L-MN for 3.2x3.2x10mm$^3$ booster in HFR, 1-port config</td>
<td>98</td>
</tr>
<tr>
<td>B.10</td>
<td>First ladder MN for 3.2x3.2x20mm$^3$ booster in LTE700, 1-port config</td>
<td>98</td>
</tr>
<tr>
<td>B.11</td>
<td>Second ladder MN for 3.2x3.2x20mm$^3$ booster in LTE700, 1-port config</td>
<td>99</td>
</tr>
<tr>
<td>B.12</td>
<td>BB-MN for bar-shaped booster in LFR, 3-port config</td>
<td>99</td>
</tr>
<tr>
<td>B.13</td>
<td>T-MN for bar-shaped booster in HFR, 3-port config</td>
<td>100</td>
</tr>
<tr>
<td>B.14</td>
<td>Ladder MN for bar-shaped booster in LTE700, 3-port config</td>
<td>100</td>
</tr>
</tbody>
</table>
Chapter 1

Introduction

In the last decade, wireless devices used by people have experienced a dramatical evolution in terms of processing capability, multimedia functionality and size. Users clearly demand that handsets include powerful hardware for their daily usage. For them it is also important the size of the wireless devices in two main senses: the apparatus shall be comfortable to use and interact with, and it shall not be bulky or heavy either due to apparent portability reasons. Obviously, there is a wide range of apparatuses, which consequently sets a different criterion for assessing all those factors. For instance, the weight of a laptop and a mobile phone are not comparable, nor is their portability comparable even though both are wireless. Still, laptops and mobile phones—and tablets, and smartwatches, etc.—have suffered a drastic reduction in weight and size.

Generally speaking, it is safe to say that all kinds of wireless devices have been made lighter, thinner and more powerful in terms of computational and graphical processing, albeit they are not strictly smaller. With the apparition of the term smart that is linked to such devices, the mainstream increase in size of the screen has, unavoidably, made the whole body and framing of the apparatus bigger yet, for example, a modern cell phone is thinner than a cellular phone from several years back and, in most cases, lighter.

The companies that manufacture any apparatus with these characteristics require of antenna engineers that design the antennas that are to be included within the encasing of such devices. In fact, nowadays, most of the antennas in a phone, for example, are custom-designed. This means that antenna engineers are given a set of specifications and requirements that the antennas being designed must comply with. Among these specifications and requirements are the space available for the antenna, the surrounding components that may detune the antenna, and the frequencies in which the antenna must be operable, to name a few. In the case of a smartphone, the number of antennas may be even greater than five, and for each of these antennas the procedure is the same.

Designing an antenna is both laborious and expensive as it is an iterative process; once
the design is ready, the antenna must be manufactured, placed inside the encasing and tested. If it does not operate well enough, the design must be modified and the whole process be repeated. Furthermore, an additional expenditure the companies must face due to custom-designed antennas is the manpower necessary for mounting and securing the antennas within the encasing of the apparatus; usually customized antennas cannot be mounted by an automated machine and, thus, personnel is required for carrying out this task.

It is possible to imagine how complex and expensive it results to make such custom designs when antennas are to be made operable for specific cellular standards, including multiband antennas which may function in different frequency ranges of the electromagnetic spectrum. Particular frequency bands and ranges are, in turn, linked to or owned by particular cellular operators, which require phones to work in their spectra and, as such, antennas shall be prepared to work in those.

Many of the problems and costs due to the use of customized antennas are alleviated or even disappear with the booster technology. This technology provides a dramatic reduction in new phone development as ground plane boosters are off-the-shelf products that, together with a matching network, may provide communication capabilities to a device. Certainly it is still necessary to design the matching networks, but that is also true for custom-designed antennas. Furthermore, not only boosters are a standard solution, as opposed to being custom-designed, but they are also convenient for pick-and-place manufacturing processes, hence that the manpower is substantially minor with respect to the traditional antennas being currently included in wireless devices. Thus, with the booster technology, providing a handset with communication capabilities becomes a cheaper process and also provides a faster device time to market.

Ground plane boosters have already proved to be a successful technology owing to their performance in current cellular frequencies. A first product based on the booster technology has been already developed and launched into the market [1].

This thesis, which has been developed at Fractus® (whose headquarters are shown in Fig. 1.1), focuses on expanding on the ground plane booster antenna technology (also called Virtual Antenna Technology™) with possible new form factors, so that, should they prove to be feasible and valid, they may finally be developed and brought onto the market. As a result, customers would be able to choose (may end up choosing) the boosters that better suits their devices based on their specific needs.

1.1. Objectives and Scope

In this thesis, ground plane booster technology is studied in pursuit of developing a demonstrator for a form factor characterized by a reduced height.
1.2. Structure of the Thesis

The main objective is to study the parameters defining the dimensions of ground plane boosters and how much they affect the impedance bandwidth of the radiating structure. The knowledge of the implications of said parameters is deemed to be fundamental for designing a booster with a form factor convenient for thin handsets with very demanding RF (radiofrequency) capabilities.

As briefly introduced, boosters are considered a successful technology thanks to an already developed product, made by Fractus®, that fully functions within the bands of interest of the present thesis, which are those of cellular services. Moreover, the existence of this product is convenient for setting a target of the new booster in terms of RF performance.

Finally, the matching circuits necessary to make the boosters operable are also part of the objective, as a couple of matching network designing options are explored, and they both condition the topology and components of the circuits that will ultimately define the performance of the boosters.

1.2. Structure of the Thesis

This thesis is organized in six chapters devoted to the study and development of a new form factor for ground plane boosters.

The first chapter, the current one, describes the objectives and scope of the present thesis and the methodology followed to accomplish the proposed objectives.

The second chapter provides an overview of main basic concepts related to antennas and ground plane boosters that will be used throughout the thesis.

In the third chapter, the ground plane booster technology is reviewed and briefly explained from a theoretical point of view for understanding what the booster technology consists in. Moreover, an exemplary radiating structure comprising ground plane boosters
from the literature is shown in terms of performance.

In chapter four, the main matching network topologies used with boosters are revised. The distinction between mismatch loss and transducer power gain based designs is provided too for a possible matching network improvement of radiating structures including boosters.

The core of the thesis is found in chapter five. The studies of all the parameters involving the design of a new form factor complying with the objective are explained, performed and tested. The performance of the proposed boosters is first simulated; prototypes are then built and their performance measured and compared to the corresponding simulations.

Finally, in chapter six, the conclusions of the thesis are presented together with some future lines of work.

1.3. Work Methodology

The methodology followed for carrying out the objectives of this thesis may be separated in three main groups of tasks: documentation, simulations, and prototyping and measuring. These groups of tasks are now described in the following subsections.

1.3.1. Documentation

It is known that research on booster technology and matching networks, the two main parts of this project, has been done prior to this thesis. Research in these fields concludes with insights and explanations which may be useful for developing a new form factor. For this reason, it is deemed necessary to collect and study papers and other bibliographical references related to these topics.

Besides the common approach of directly searching for said kind of documents in scientific digital databases or publications such as IEEE Xplore®, Microwave and Optical Technology Letters or International Journal of Microwave Science and Technology, to name a few, the search for bibliography is also performed using the ZyLAB® software available at Fractus®.

ZyLAB®, a software convenient for knowledge management systems, generates and maintains a database of documents introduced by the user. At any point the database may be queried so as to recover documents that match a search string. Interestingly, other than traditional fields or tags such as authorship, publication date, or journal where it was published, the system also looks up for matches of the search string in the content of the documents, so the text of the papers is also inspected. Thus, thanks to ZyLAB®, it is possible to look for relevant bibliography within a database of papers which, in the case of Fractus®, knowledge management system, is very large.
1.3. Work Methodology

1.3.2. Simulations

One important part of this thesis is simulating radiating structures that comprise ground plane boosters according to the proposed form factors. Owing to the results of said simulations, it is possible to anticipate the performance of different boosters and, therefore, alleviate the amount of prototypes necessary to find a proper solution.

For this reason, first a radiating structure must be digitally modeled in 3D. This is done with Mentor Graphics® IE3D™, a software that supports designing models for microwave engineering, characterizing the materials as dielectric layers and conductive layers with their respective characteristic parameters, and simulate their performance using the method of moments. The output of the simulations may be, for example, the S-parameter matrix that characterizes the modeled structure for a range of frequencies. A screenshot taken from Mentor Graphics® IE3D™ is shown in Fig. 1.2.

![Mentor Graphics® IE3D™](image)

Figure 1.2: Mentor Graphics® IE3D™

It is necessary to study the performance of the modeled structure with a matching network too. In spite of the existence in IE3D™ of a functionality for adding a matching circuit and recompute the S-parameter matrix, this task is performed using National Instruments™ AWR Design Environment™ - Microwave Office®.

Microwave Office® (a screenshot from the software is shown in Fig. 1.3) is oriented to the design of matching networks with ease and straightforward optimization based on several criteria established by the user. So, the files containing the S parameters simulated by IE3D™ are imported to Microwave Office® and, then, matching networks are added for a complete radiating structure simulation. The same software allows representing, among others, the S parameters in rectangular and Smith charts, which is helpful for assessing the components that have to be added or modified so as to improve the response of the
full radiating structure. Both types of graphs are included in this thesis, when deemed necessary, presenting the results of the simulations.

![Figure 1.3: Microwave Office®](image)

The results obtained from the simulations of the 3D models together with the matching circuits are a rough estimation of the real performance of the whole radiating structure guiding towards the form factor that, ultimately, provides the best performance.

### 1.3.3. Prototyping and Measuring

Once all the simulations have been run and the results are considered to be satisfactory, the simulated radiating structure and its matching circuits have to be built in a real prototype. Afterwards, their S parameters and antenna efficiency have to be both measured.

For prototyping, the layout of the printed circuit board, commonly referred to as PCB, is drawn in *AutoDesk® AutoCAD®*. With the layout printed, the PCB is made in *Fractus®* chemical laboratory (Fig. 1.4) using a 1mm-thick FR4 material.

![Figure 1.4: Photo of Fractus® chemical laboratory](image)
For measuring the RF performance of the PCBs, two equipments have been used. Agilent Technologies E5071B network analyzer (Fig. 1.5) has been mainly used to measure the S parameters. This network analyzer is capable of measuring from 300kHz up to 8.5GHz. An anechoic chamber including a Satimo SG 32 shown in Fig. 1.5 has been the equipment in charge of measuring the prototypes’ antenna efficiency and radiation patterns. The final prototypes presented herein include information regarding the antenna efficiency by means of rectangular graphs with the corresponding reflection coefficient measured in the network analyzer.

Figure 1.5: (left) Photo of Agilent Technologies E5071B network analyzer; (right) Photo of Satimo SG 32 anechoic chamber

1.4. Patents

Resulting from the inventive work done in this thesis, a new patent application has been produced and filed to several patent and trademark offices [2][3].
Chapter 2

Basic Concepts Related to Antennas

Throughout this thesis, several concepts and parameters related to RF systems and antenna theory in general, are used for explaining and defining the operation of antennas and radiating structures according to this project. Certain knowledge in microwave engineering is assumed reader-side and, thus, only few concepts are described herein.

2.1. Reflection Coefficient

The voltage reflection coefficient, commonly referred to as reflection coefficient, is a complex magnitude that relates the amplitude of the reflected voltage wave normalized to the amplitude of the incident voltage wave for a specific point of a transmission line

\[
\Gamma = \frac{V_0^-}{V_0^+} = \frac{Z_L - Z_0}{Z_L + Z_0}
\]  

(2.1)

Wherein \(\Gamma\) stands for voltage reflection coefficient; \(V_0^-\) and \(V_0^+\) stand for the amplitude of the reflected voltage and the amplitude of the incident voltage waves, respectively; \(Z_L\) stands for the input impedance of the load; and \(Z_0\) stands for the characteristic impedance of the transmission line.

In the particular case in which the load impedance \(Z_L\) is equal to the characteristic impedance of the transmission line \(Z_0\), the load is said to be matched to the line since there is no reflection of the incident wave, resulting in a reflection coefficient \(\Gamma = 0\).

The standing wave ratio (SWR), sometimes known as voltage standing wave ratio or VSWR, is a measure of the mismatch of a line defined as in (2.2) just considering the magnitude of the reflection coefficient.

\[
\text{SWR} = \frac{|\frac{V_0^-}{V_0^+}|}{|\frac{V_0^-}{V_0^+}| + |\frac{V_0^+}{V_0^-}|} = |\frac{Z_L - Z_0}{Z_L + Z_0}|
\]  

(2.2)
Chapter 2. Basic Concepts Related to Antennas

\[ SWR \equiv S = \frac{1 + |\Gamma|}{1 - |\Gamma|} \tag{2.2} \]

With the reflection coefficient and/or the standing wave ratio it is possible to define further parameters such as the antenna bandwidth.

2.2. Antenna Bandwidth

Any antenna structure configured to transmit and/or receive electromagnetic wave signals is designed to do it effectively in a limited range of frequencies or, equivalently, wavelengths. The ranges in which certain parameter or parameters of the antenna are within specific preestablished limits are known as antenna bandwidth [5].

One of the most typical ways to characterize the antenna bandwidth is as the range of frequencies in which the reflection coefficient or the voltage standing wave ratio, both parameters related to each other and dependent on the input impedance of the antenna, do not exceed a threshold. In this case, when a frequency is included within the bandwidth of an antenna, then the antenna is said to be matched in impedance at that frequency.

The antenna bandwidth might be expressed either in frequency units when the difference between its upper and lower frequencies is computed, or as a percent when such difference is divided by the central frequency of the bandwidth. Both of these values may be used interchangeably as they are directly related. If any information is lacking, it affects both values, as the limiting frequencies cannot be inferred based on any of the two previous definitions.

2.3. Quality Factor Q of Antennas and Inherent Bandwidth

As explained in the previous section, the bandwidth of an antenna is a key parameter for characterizing the range of frequencies in which the antenna has a good performance. In [6] an approximated expression for the inherent bandwidth of a tuned antenna has been derived based on the antenna’s input impedance.

To retrieve and compute the bandwidth of an antenna, first it is necessary to compute an approximated quality factor Q of the antenna, a parameter fundamentally defined by the fields of the antenna (2.3). This parameter describes the overall performance of an antenna and the inherent physical limitations regarding the size of the antenna. The larger the Q factor is, the larger the amounts of reactive energy that are stored in the near zone field, which implies having large currents, high ohmic losses, narrow bandwidth and a large frequency sensitivity [7]. The lower the Q is, the more broad band performance the antenna features.
2.3. Quality Factor Q of Antennas and Inherent Bandwidth

\[ Q(\omega_0) = \frac{\omega_0 W}{P} \]  

(2.3)

Wherein \( \omega_0 \) stands for the particular value of angular frequency in which the quality factor is being computed; \( W \) stands for the reactive energy in the antenna; and \( P \) stands for the power dissipated in the network.

This Q factor may be expressed in function of the input impedance, and is inversely proportional to the bandwidth of the antenna. An approximated mathematical definition of the Q factor in terms of the antenna’s input impedance, defined as \( Q_Z \), is shown in (2.4).

\[ Q_Z(\omega_0) = \frac{\omega_0}{2R(\omega_0)} |Z'(\omega_0)| = \frac{\omega_0}{2R(\omega_0)} \sqrt{\left(\frac{dR(\omega_0)}{d\omega_0}\right)^2 + \left(\frac{dX(\omega_0)}{d\omega_0}\right)^2 + \left(\frac{X(\omega_0)}{\omega_0}\right)^2} \]  

(2.4)

Wherein \( R(\omega_0) \) stands for the resistance of the input impedance at \( \omega_0 \); and \( X(\omega_0) \) stands for the reactance of the input impedance at \( \omega_0 \).

The value resulting from the computation of the Q factor using the previous equation is, according to [6], quite accurate for antennas modeled as series RLC circuits—like ground plane boosters, although boosters are not antennas—in resonant and antiresonant frequency ranges; in the numerical simulations run during this thesis, the Q factor computed for boosters seems to validate the assertion that the results are accurate for series RLC circuits, even when what is studied is not an antenna in a literal sense.

For the record, the only situation in which there were inaccuracies in the computed bandwidth occurred for antennas resembling parallel RLC circuits which, at some frequency, had an imaginary part of the impedance equal to zero. Adding a series inductance or capacitance to the impedance such that the imaginary part was not equal to zero for any frequency, said inaccuracy was avoided and the resulting bandwidth was again quite precise. This case, however, is not relevant in the development of a new form factor for ground plane boosters as boosters do not behave as parallel RLC circuits.

As mentioned above, the \( Q_Z(\omega_0) \) is inversely proportional to the bandwidth, which throughout this thesis is referred to as the inherent bandwidth because it is strictly related to the antenna’s input impedance. The standing wave ratio as defined in (2.2) is necessary for calculating the inherent bandwidth figure since it depends on the SWR threshold required for the antenna to perform correctly. So, with \( Q_Z \) and the SWR \( (S \) in the equations), the inherent bandwidth may be defined as (2.5).

\[ BW_Z(\omega_0) = \frac{f_2 - f_1}{f_0} = \frac{1}{Q_Z(\omega_0)} \frac{S - 1}{\sqrt{S}} \]  

(2.5)

Making the computation of \( BW_Z(\omega_0) \) gives as a result a fractional bandwidth, namely
a ratio between the difference of the upper \( f_2 \) and lower \( f_1 \) frequencies whose standing-wave ratio is equal to the one established in the equation, over the central frequency \( f_0 \) for that particular bandwidth.

If the inherent bandwidth is to be calculated using the \( Q_Z \) expression from (2.4) in a bandwidth wide enough that exhibits multiple resonances, the result has been proved in [8] also to be consistent long as the SWR limit set is substantially small. Albeit no multiple resonances occur in the cases studied for this thesis, the maximum permitted SWR—for considering a proper matching—has been defined to be 3, which is equivalent to a reflection coefficient being less than -6dB.

2.4. Antenna and Radiation Efficiencies

The magnitude that ultimately measures the performance of an antenna structure in terms of wave transmission and reception is the antenna efficiency \( \eta_a \). Said efficiency denotes how effectively the power delivered to the system—the whole antenna structure—is finally radiated by the antenna (2.6).

\[
\eta_a = \frac{P_{rad}}{P_{sup}} = \frac{P_{rad}}{P_{rad} + P_{loss,AS}}
\]  

(2.6)

Wherein \( P_{rad} \) stands for power radiated by the antenna structure; \( P_{sup} \) stands for power supplied by the generator to the antenna structure; and \( P_{loss} \) stands for power losses.

Part of supplied power is dissipated as heat by different elements forming the antenna structure, such as the components from the matching network, which are lossy; the dielectric, that is also lossy; and the antenna, the impedance of which is formed by a radiating part and a lossy part as well.

A different parameter that is related to the antenna efficiency is the radiation efficiency \( \eta_r \). In contrast to the antenna efficiency, the radiation efficiency represents the ratio between the power that is radiated into space by the antenna and the power that is actually delivered to the antenna itself (2.7).

\[
\eta_r = \frac{P_{rad}}{P_{del}} = \frac{P_{rad}}{P_{rad} + P_{loss,Ant}}
\]  

(2.7)

Wherein \( P_{del} \) and \( P_{loss,Ant} \) respectively stand for the power delivered to the antenna and the power dissipated as heat by the antenna.

Therefore, as the power is directly delivered to the antenna, it can be simplified as the ratio between the radiation resistance of the antenna and the total resistance of the antenna, which includes the radiation resistance and the losses resistance. How well or bad the antenna is matched to the actual generator is not considered because that is already
taken into account in the antenna efficiency. A good approximated relation between the two efficiencies is shown in (2.8) where the reflection coefficient (represented as the S parameter \( S_{11} \)) considers the effect of the impedance matching.

\[
\eta_a \approx \eta_r \left(1 - |S_{11}|^2\right)
\]  
(2.8)

For an exact relationship between the antenna and the radiation efficiency, not only the matching has to be considered but also the power that is lost within the matching network [9]. An accurate relationship between the two efficiencies is expressed in (2.9) [10].

\[
\eta_a = \frac{P_{rad}}{P_{sup}} = \eta_m \eta_r
\]  
(2.9)

Wherein \( \eta_m \) stands for matching efficiency.

This efficiency assesses the matching quality and the amount of losses that the matching circuit features, however both parts are dependent on the terminal impedance of the antenna and, thus, it is not computed in a straightforward manner. As such, equation (2.9) is rather uncommon and its use is merely confined to the literature.

For quantifying the amount of power that is lost in the matching circuit, the quality factor Q, this time for lumped components, is used. Said factor is now explained in the following section.

### 2.5. Quality Factor Q of Lumped Components

Lumped components used in matching networks are not ideal. Usually, capacitors and inductors are used in a matching network, and not only their associated impedance is dependent on the frequency, but they are lossy as well, meaning that a parasitic resistance is present in the components.

The parasitic resistance is normally represented with two different parameters: the quality factor Q of the component, and the equivalent series resistance or ESR. Both are frequency-dependent and are useful for assessing the amount of power dissipated as current flows through the components.

The Q factor is generally a rather big—compared to the ESR—natural number that is equivalent to the equation (2.10).

\[
Q = \frac{|X_g|}{R_g} = \frac{2\omega U}{P}
\]  
(2.10)

Wherein \( X_g \) stands for the reactance of the component; \( R_g \) stands for the (parasitic) resistance of the component; \( \omega \) stands for the angular frequency; \( U \) stands for the average
energy stored in the component; and $P$ stands for the average power dissipated in the component [10].

The larger the $Q$ factor is, the less lossy the component is as the associated parasitic resistance is smaller with respect to the reactance. Albeit the reactance of the components depends on the frequency, so does the parasitic resistance. For this reason, the $Q$ factor is represented in the specifications or datasheets of the components as a graph, such as the ones in Fig. 2.1, where the curves for the $Q$ factor of an SMD capacitor and an SMD inductor are represented.

Figure 2.1: (left) $Q$ factor versus frequency of a 0402 SMD High-$Q$ 5pF-capacitor from Murata; (right) $Q$ factor versus frequency of a 0402 SMD High-$Q$ 5nH-inductor from Murata

In contrast, the ESR directly represents the value of the parasitic resistance in ohms. Rectangular charts analogous to the $Q$ factor are represented in Fig. 2.2 for illustrative purposes only. Said parameter is helpful for quickly assessing how much power may be dissipated in the matching network in a per-component basis, however for assessing how lossy is the matching network overall the $Q$ factor is much more convenient as the quality of the components may be easily compared. Generally, the quality of inductors is expressed in terms of their $Q$ factor, whereas the quality of the capacitors is expressed in terms of the ESR parameter.

Figure 2.2: (left) ESR versus frequency of a 0402 SMD High-$Q$ 5pF-capacitor from Murata; (right) ESR versus frequency of a 0402 SMD High-$Q$ 5nH-inductor from Murata
Chapter 3

Ground Plane Boosters

3.1. Introduction

Since the conception of the first mobile phones, the antennas enabling voice communications have undergone big transformations. The most apparent change has been the suppression of the external antenna as small internal antennas were developed which, eventually, have become the current standard in phones, and also in other types of handsets and portable devices.

The antennas had to adapt to the size constraints as new cellular standards were being developed and adopted. This posed a great challenge to antenna engineers as the antennas had to be smaller and operable at different frequencies; the trend in adding antennas exclusively dedicated to a single frequency band each had to be replaced by the reuse of an antenna for many cellular standards; with this goal, the multilevel antennae [11] were developed. Since then, usually every antenna for cellular and communication services has its design customized to the handset in which it is to be allocated; the task of making custom antennas being complex, laborious and, obviously, having an impact on the cost.

It is not trivial to design an antenna that is both efficient and small: normally radiating elements are resonant and provide operation at the frequencies at which the physical dimensions of the antenna are one fourth or, in the case of inverted-F antenna structures, one eighth of the wavelength. A quite revealing numerical example is the operation in 2G standards such as GSM850 or GSM900 that work around 900MHz. At this frequency, the wavelength in free-space is about 33 centimeters, thus a fourth being about 8 centimeters. Additionally, the very same antenna may have to operate a 3G standard like PCS1800 at about 1.8GHz, among other services (e.g. 2G, 3G and 4G standards working at different frequencies). Different current antenna structures and designs are depicted in [12], with particular chapters focused on monopoles and planar antennas including inverted-F topologies.
Traditionally, the geometry of the antennas has been exploited so they adapted to the particular set of constraints the engineers were facing, and not so much attention was paid to other important parts of an antenna structure such as the ground plane. Recently, more interest has been put in the design and use of the ground plane in an attempt to improve the antenna structure in terms of its overall performance [13]-[18].

Ground planes have been taken a step further and, in pursuit of standardizing the way antennas are conceived, ground plane boosters have been researched and developed in the last few years [19]-[26]. With this technology it is possible to have devices without antennas in the literal sense and still be able to establish data and voice communications. The use of ground plane boosters, thus, essentially makes the devices antennaless.

This chapter is structured in two parts. First, a section devoted to explaining the theory basics of ground plane boosters is presented. The subsequent sections briefly show an exemplary radiating structure from the radioelectric performance point of view, being this part further expanded in chapter 5 alongside with the performance of the new researched boosters.

3.2. Theory

An essential part included in any electronic device is the ground plane, a conductive element with low electric potential. Thanks to the ground plane, electrons may generate an electric current that feeds the distinct components and chips in the device. In addition, a ground plane also allows an antenna to function due to the potential difference present in the antenna structure, so it is a very important part of it that further contributes to the radiation and reception of electromagnetic waves. Antennas, being resonant, are prone to capture and transmit electromagnetic waves with wavelengths determined by the dimensions of the antenna, and make the physical reception and radiation of the waves together with the ground plane.

The research in ground plane boosters focused on the characteristics of a ground plane which, normally, features large dimensions making it highly convenient for becoming the main radiator of a device and, thus, act as an antenna. However, the ground plane alone does not present an electrical dimension suitable for capturing and transmitting electromagnetic waves as opposed to an antenna, which has a conductor already designed for being excited at one or more specific frequencies or wavelengths. For this reason, it is necessary to somehow excite the ground plane enabling it as the radiator of the device.

Aided by the characteristic modes theory, it is possible to study and assess how the ground plane contributes to the radiating process; said theory, in turn, allows systematical analysis and design of antennas for wireless devices. The characteristic modes theory describes the natural modes characterizing the radiating structure as electrical current
3.2. Theory

eigenfunctions that are related to the shape and size of the conducting object [27].

When the conducting elements forming the radiating structures are defined in terms of form and dimensions, the radiation modes may be computed. The eigenvalue $\lambda_n$ of these modes provides valuable information for optimally exciting the radiation mode sought: when the eigenvalue is positive, the mode contributes to store magnetic energy, whereas if it is negative then the contribution is to store electric energy. In addition, the smaller the eigenvalue is, the higher the contribution of the mode to the total surface current density is, wherein an eigenvalue equal to zero means that the mode is in resonance [28]. A graph representing the eigenvalues and the modal significances for a range of frequencies is shown in Fig. 3.1, in which there is featured a rectangularly-shaped ground plane.

![Graph showing eigenvalues and modal significances](image)

**Figure 3.1:** Eigenvalues and modal significances for first and second predominant modes $\lambda_1$ and $\lambda_2$ for a rectangularly-shaped ground plane sized 100x40mm$^2$; the ground plane representation corresponding to the current distribution for the first radiation mode at 892MHz [27]

Such ground plane, for a typical mobile phone size (on the order of 100x40mm$^2$), is characterized by a predominant longitudinal mode as seen at the two longer edges in the aforementioned figure. It can be further appreciated that the modal significance of $\lambda_1$ has a much larger magnitude than $\lambda_2$ in the whole frequency range.

These longitudinal modes can be advantageously excited for proper electromagnetic wave radiation and reception and, thus, avoid the use of antennas. When antennas are included in a radiating structure, their radiation modes are excited owing to their resonance and thus the natural modes of the ground plane are not exploited.

So, for said ground plane excitation, ground plane radiation boosters are used. The
booster technology consists in rather small-sized elements designed with the sole purpose of effectively coupling energy to the ground plane, so the ground plane then advantageously uses its own radiation modes for the radiation process. Owing to their reduced size, the boosters do not contribute to the radiation of electromagnetic waves—even though certain physically inevitable radiation occurs, it is remotely small compared to the radiation of the ground plane.

Boosters may either excite modes whose eigenvalue is positive, namely store magnetic energy, or negative, which store electric energy. So, depending on the fields to be excited, electric or magnetic, either the boosters are physical conducting elements or indentations in the ground plane are added like the magnetic boosters from [29]. Judging by the modes shown in Fig. 3.1, the maximum of the current distribution is in the middle of the edge, and its minimums are at the limits of the edge, a fact that has to be considered for placing the correct boosters in the optimal positions. For instance, the radiation boosters in Fig. 3.2 are two cube-shaped boosters placed in the corners exciting the longitudinal modes of the ground plane in two different frequency regions as in [20], although for less-demanding devices, one single radiation booster for all necessary communication services may suffice [19].

![Figure 3.2: Rectangularly-shaped ground plane with two cube-shaped boosters located in the corners for independent frequency region excitation [27]](image)

Thanks to the proper addition of radiation boosters, the radiation modes of the ground plane are effectively excited and thus no antenna is necessary for the device to be capable of providing operation in cellular or another services. Yet with the radiating system as it is, the bandwidth requirements are not fulfilled as the $S_{11}$ will not be low enough within the whole frequency ranges of interest.
To alleviate this situation, a matching network that adjusts the antenna’s input impedance so as to provide an impedance match is needed. Even though the particularities of the matching networks in structures including boosters are covered in chapter 4, minor details are provided in the following section regarding the performance of ground plane boosters as, among the results it presents, there are some complete radiating systems (with boosters including matching networks) as well.

3.3. Performance

Ground plane radiation boosters may have different shapes and sizes; however, the question is whether particular boosters positioned somewhere in a printed circuit board are capable of exciting the radiation modes of a ground plane in an effective manner.

A compact booster that has already been developed is one with a cubic form factor with a volume of 125mm$^3$, like the ones shown in Fig. 3.2. It has been verified that its performance when used with a ground plane with the typical size of a mobile phone, together with a matching network, is satisfactory according to the standards in the cellular industry.

As what is sought is operation in at least two separate frequency regions, it is convenient to use two boosters: one per frequency region. The frequency regions comprise the 824MHz to 960MHz frequency range, and the 1.71GHz to 2.17GHz frequency range; covering both ranges entirely, pentaband operation is ensured. In these, according to the graph from Fig. 3.1, one booster will be exciting the predominant mode $\lambda_1$ not being in resonance (around the 900MHz), whereas the other booster will be exciting the predominant mode $\lambda_1$ being much closer to the resonance (the eigenvalue is almost zero within the whole frequency range). Again, this happens for a ground plane with a rectangular shape and the longer and shorter edges being 100mm and 40mm, respectively. For a different ground plane, the characteristic modes theory allows to compute of its radiation modes.

So, in a structure such as in Fig. 3.2, two feeding ports are added at which the electromagnetic wave signals from one or the other frequency region are both injected and retrieved, depending if the device is transmitting or receiving. Between the booster and the feeding port, the matching networks are installed.

The task of the matching networks—providing impedance matching—is to move the input impedance to the center of the Smith chart, as much as possible, for the range of frequencies of interest. So, considering Fig. 3.3, each of the two matching networks should translate the part of the curve delimited by its frequency limits to the center, wherein the impedance of the radiating structure clearly has the behavior of a series RLC circuit.

As the impedance, in this case, is clearly capacitive—in the lower half of the Smith chart,—it is necessary to shift it upwards with an inductor. For this reason, the matching
Figure 3.3: Smith chart showing the simulated input impedance of a radiating structure comprising a rectangular ground plane (100x40mm$^2$) with cube-shaped ground plane radiation boosters (5x5x5mm$^3$) located at the corners; markers show the limiting frequencies for the two frequency regions [20].

networks for both boosters include a series inductor as the first element of the circuit, making their respective reactances closer to zero, so the inductor necessary for the lower frequency range is substantially larger than for the higher frequency range. Afterwards, the matching circuits are completed with combinations of capacitors and inductors—the shunted LC resonators are explained in the following chapter—as represented in Fig. 3.4, helping to create respective impedance loops contained within the SWR=3 region (see Fig. 3.5).

Figure 3.4: (left) Matching network for the 824-960MHz frequency range; (right) Matching network for the 1710-2170MHz frequency range [27].

These impedance loops ultimately provide the antenna bandwidth in each frequency region as seen in Fig. 3.6.

Once a prototype according to this exemplary radiating structure was built, the two ports were combined into a single one using a transmission line and a junction. Notch filters were added to both matching circuits for low and high frequency filtering that would not cause problems in the feeding; and after adjusting the lumped components to
Figure 3.5: (left) Smith chart showing the simulated input impedance of the radiating structure from Fig. 3.3 with a matching network for the 824-960MHz frequency range; (right) Smith chart showing the simulated input impedance of the radiating structure from Fig. 3.3 with a matching network for the 1710-2170MHz frequency range [27]

Figure 3.6: Rectangular chart showing the simulated $S_{11}$ parameter of the radiating structure from Fig. 3.3 using matching networks [27]
commercially available values, the prototype was measured in its only port both in terms of $S_{11}$, represented in Fig. 3.7, and efficiency, represented in Fig. 3.8.

![Figure 3.7: Rectangular chart showing the measured $S_{11}$ parameter of the radiating structure from Fig. 3.3 using matching networks with notch filters [27]](image)

![Figure 3.8: Rectangular chart showing the measured antenna efficiency $\eta_a$ and computed radiation efficiency $\eta_r$ of the radiating structure from Fig. 3.3 using matching networks with notch filters [27]](image)

The radiating system features good reflection coefficient and efficiency both in the low and high frequency regions, using just two ground plane boosters with a size of $5\times5\times5\text{mm}^3$ each, totaling a volume of $250\text{mm}^3$. The radiation patterns are quite omnidirectional as represented in [27] which is convenient for cellular communications. This example demonstrates that it is possible to use the natural modes of a ground plane for acting as an antenna, although the modes shall be correctly excited as ground plane boosters precisely do.

This section has shown an exemplary radiating structure using the booster technology. In the following chapters, a different radiating structure will be considered for the design of a booster with a new form factor.
Chapter 4

Matching Networks for Ground Plane Boosters

4.1. Introduction

As stated in chapter 3, ground plane boosters are small-sized elements that couple energy from or to the ground plane, exciting said ground plane that, in turn, may act as the main radiator for the whole radiating system. Other than said excitation, impedance matching must be ensured for the correct operation of the device in the bands of interest.

Generally, when using resonant antennas, the antenna bandwidth is already available and minor alterations to the input impedance have to be performed via lumped or distributed elements forming the matching network. Adding a matching network does not alter the radiating elements and it may compensate for the rapid frequency variations of the input impedance of an antenna [30].

In the context of this project, the frequencies of interest range approximately from 0.7GHz up to 2.7GHz. At these frequencies, cubic radiation boosters with dimensions of $5\times5\times5\text{mm}^3$ feature a negative reactance. Said boosters, actually, are non-resonant as the frequency at which the imaginary part of the impedance equals zero is above 3GHz.

However, in the case of ground plane boosters, most of the frequencies of interest do not form part of the antenna bandwidth: they do not comply with the $S_{11} \leq -6\text{dB}$ or $\text{SWR} \leq 3$ criteria. If power is to be injected at these frequencies, most of it would be reflected and thus not radiated into free space.

Additionally, the antenna bandwidth necessary for the three frequency regions or bands forces the radiating system to be broadband at each one of them. To put up some numbers, the LTE700 band, with a center frequency of 748MHz, has a bandwidth slightly larger than 13%; the low frequency region, with a center frequency around 900MHz, demands a system
with a 15% bandwidth; and, finally, the high frequency region which has a center frequency of 2.2GHz, requires a bandwidth of almost 45%. Although the bandwidth figure for the high frequency region is quite large, the 13% and 15% shall not be considered narrow as having a long wavelength they are, in fact, very demanding for a radiating system.

A radiating system as disclosed in this thesis, when mismatched, may achieve antenna efficiencies of 10% or even less in the frequencies of interest. For this reason, it is necessary to match the radiating system for the frequencies that should be operated and, thus, ensure effective wave transmission and reception.

When developing a matching network, a very low reflection coefficient is sought so the power does not get reflected back to the generator. Nevertheless, it is also important to match the radiating structure in a manner such that minimum power is dissipated in the matching network, which inherently has losses due to parasitic resistances in the components.

Since the least lossy network is not the one that provides the best impedance matching, two different approaches are possible in the engineering of matching networks: either the reflection coefficient is optimized (as in mismatch loss-based matching networks;) or the transducer power gain (TPG) is optimized.

4.2. Mismatch Loss Based Designs

The first approach for designing a matching network consists in minimizing the amount of signal reflected to the port in which such signal is injected or fed.

In a mismatch loss based design, in addition to the apparent advantage of having more power delivered to the rest of the system—the power available at the generator is the difference between the power delivered and reflected at its terminals—such minimization is also convenient for avoiding misbehaviors in different generator circuitry components due to currents originated because of power reflections.

If the components were lossless, this approach for designing matching networks would make the radiating system provide the best possible performance as, then, the equalities from (4.1) would hold [31].

\[
|S_{21}|^2 = 1 - |S_{11}|^2 = 1 - |S_{22}|^2 \quad (4.1)
\]

As the transducer power gain \( |S_{21}|^2 \) just depends on the reflection coefficient on either port \( S_{11} \) and \( S_{22} \), but only one at a time, minimizing any of the reflection coefficients as done in mismatch loss based designs would deliver the maximum power. In the end, the transducer power gain is the parameter that determines the amount of power that is supplied to the radiating element.
4.3. Transducer Power Gain Based Designs

From a practical point of view, considering losses or not, a mismatch loss based circuit may be designed rather straightforward using the Smith chart. The effect of capacitors and inductors put in series or in parallel is known beforehand, so the engineer designing the circuit may tailor the topology makes the impedance closer to the center, namely minimizing the reflection coefficient $\Gamma$, with the least amount of components possible. It is also a matter of the engineer's expertise to assess which components should be prioritized based on quality factors, effective inductance and capacitance values, and tolerances. Although the focus is put on providing good impedance matching, it is convenient to already design the circuit minimizing deviations and power losses. Additionally, once a matching circuit is synthesized in a simulation software like *Microwave Office®,* components may be modified on-the-go when being mounted on a prototype thanks to a network analyzer.

4.3. Transducer Power Gain Based Designs

Although commercially available capacitors and inductors feature a high quality factor $Q$, the losses the radiating system incurs in are not negligible. Therefore, it is necessary to consider the parasitic resistances in the components used as they will affect more or less notably the performance of the radiating system.

Whether power is being reflected back to the generator or dissipated in the form of heat in the matching network's components, the transducer power gain approach still focuses on maximizing the amount of power that is transferred to the load.

As TPG matching networks are capable of providing more power than mismatch loss based circuits, it can be concluded that featuring a higher reflection coefficient may be less disadvantageous if the components are not so lossy.

In the case of lossless matching circuits, (4.1) is true. However, as losses kick in in the system, the equalities do not hold anymore (4.2) [31].

$$\left|S_{21}\right|^2 \neq 1 - \left|S_{11}\right|^2 \neq 1 - \left|S_{22}\right|^2$$  \hspace{1cm} (4.2)

The fact that the right-most equality does not hold true either is especially interesting; it means that the losses in the matching circuit are not the same from one port or the other, even in the case of circuits consisting of passive components.

As it turns out, a matching network may provide good impedance matching to one port, yet the power reflected sees large losses within the network, whereas looking from the other port the same network may be poorly matched but see small losses in the circuit. According to [31], it is advisable to check whether the $S_{11}$ and $S_{22}$ parameters are similar. In case they are not roughly the same it can be inferred that the matching circuit features large losses. It may be, however, rather difficult to take both measurements in a prototype.
At most, it may be checked in the simulations in a simplified manner.

An example of transducer power gain maximization not being equivalent to reflection coefficient minimization is now presented. Using a ground plane booster with size 5x5x5mm³ in a ground plane sized 120x60mm², an identical topology for a matching circuit is tested three times with different component values. The schematic from Fig. 4.1 shows in a color and symbol coded way the components used at each run; said schematic belongs to a broadband matching network for the 824-960MHz frequency range that will be described in the following section. These matching circuits are simulated both in lossy and lossless conditions.

Figure 4.1: Three matching networks used in for transducer power gain examples; a cube-shaped booster sized 5x5x5mm³ uses a single color and symbol coded matching network each time; Q and ESR parameters for central frequency indicated

With these matching networks, the impedance seen at the feeding port is represented in the Smith chart from Fig. 4.2.

The brown (triangle) matching circuit creates a big impedance loop, the blue (square) matching circuit creates a medium-sized impedance loop, and the green (circle) matching circuit creates a small impedance loop. As in the Smith chart only the relevant frequencies are represented, the loops are not appreciated; if a broader range was to be shown, the impedance loops would be drawn very clearly. In any case, from the radius of the curves the size of the loop could be inferred.

By simple inspection of the Smith chart, which is depicting the impedance for the lossless cases, the best impedance matching would be achieved by the small loop, then the medium-sized loop, and in last place, the big loop. If the S_{21} parameter is plotted in dB versus frequency, being 0dB the equivalent to delivering the same amount of power to the radiating system that is supplied by the generator, the result is as shown in Fig. 4.3.

Even though the medium and small impedance loops average a slightly better S_{21} parameter than the big-sized loop for the lossless scenario, in the lossy case it is the big-sized loop the one getting a higher average: 78 % versus 77 % and 75 % for the medium and
4.3. Transducer Power Gain Based Designs

Figure 4.2: Smith chart showing the simulated input impedance of a cube-shaped booster sized 5x5x5mm$^3$ using respective color and symbol coded lossless matching networks from Fig. 4.1.

Figure 4.3: Rectangular chart showing showing the simulated $S_{21}$ parameter of a radiating system comprising a cube-shaped booster sized 5x5x5mm$^3$ using respective color-coded lossless matching network from Fig. 4.1; dotted lines representing the lossless cases, and the solid lines representing the lossy cases. Averages of $S_{21}$ parameter in linear for lossless case: big loop 87%, medium-sized loop 88%, small loop 88%. Averages of $S_{21}$ parameter in linear for lossy case: big loop 78%, medium-sized loop 77%, small loop 75%.
small loops, respectively.

This demonstrates that when matching circuits according to the transducer power gain maximization approach are designed, a trade-off between reflection coefficient and losses must be reached. It is more complex to assess the effect of both phenomena when, simply put, moving the impedance to the center of the Smith chart is not always the goal. In the end, there is always the dilemma whether the gain in reflection coefficient achieved using one component or the other is smaller than the losses due to power dissipation and vice versa.

It is known, however, that not employing components that store energy in the same form than the load will improve the efficiency. Thus, if the load is capacitive, the radiating system will be more efficient if the synthesized matching circuit only comprises inductors, rather than using inductors and capacitors. This is because the energy stored in the inductors will be exchanged with the capacitors from the network rather than with the load [10]. Achieving a good match with just inductors is not simple though, and for the most part a combination of both types of components is used, but always trying to minimize the number of capacitors.

Finally, maximizing the TPG would be more feasible considering equation (2.9), where the optimization of the transducer power gain would imply maximizing the matching efficiency $\eta_m$, a variable that comprises both the effect of the mismatch and the losses, and weighs the radiation efficiency $\eta_r$. The matching efficiency, however, is not a parameter assessing the quality of the matching circuit as an independent entity, but rather the quality of the matching circuit considering the input impedance of the booster: the impedances of the matching network components are dependent on the impedance of the load [10].

Computing $\eta_m$ is not very practical because the bandwidths that have to be considered in this thesis are rather broad, so the impedance of the booster varies significantly within the band and the matching efficiency would have to be calculated for many frequencies. Additionally, the components also behave differently at each frequency, so many Q factors and the effective values of their reactances should be considered for accurate matching efficiencies, not to mention the tolerances of the components themselves. Therefore, optimization of the transducer power gain is done directly via S parameters using Microwave Office®.

4.4. Matching Networks for Ground Plane Boosters

Several matching network topologies are considered for providing a good impedance match to the ground plane boosters. The largest topology comprises, at most, four lumped components, and the smallest one comprises two. Some of these are now depicted as the circuits used in the simulations and prototypes in the following chapter are included in
For the LTE700 band, a 4-component ladder topology (Fig. 4.4) provides the bandwidth necessary at frequencies as low as 700MHz and 800MHz.

The series inductor has to compensate the capacitive behavior of the ground plane booster. As such, the inductor is usually on the order of 40nH, which features an ESR on the order of 3Ω to 4Ω in the High-Q 0603 SMD component. This ESR, then, is the most significant one among all the components used, and even though the rest are lossy as well, the series inductor dissipates slightly more power than the average.

With respect to the low frequency region, the center frequency of which is 892MHz, a 4-component broadband network topology is used (Fig. 4.5). It has been shown in [32][33] that the bandwidth of antennas featuring an impedance modeled by RLC both series or parallel may be improved by simple broadband matching networks.

It has been derived that the use of a broadband matching network, for an SWR equal or better than 3, may increase the antenna bandwidth up to 2.45 times [27] with respect to the inherent bandwidth. Thus, computing the inherent bandwidth, it may be estimated the achievable antenna bandwidth with a broadband matching network using a 2.45 multiplying factor.

A particularity of said matching network is the existence of a resonator in the form of a parallel LC circuit. The resonance of this circuit is usually chosen to be nearby the center frequency of the band. Even though there are many possible value combinations that cause
a resonance in the middle of the band, the election of the capacitor and inductor is not random as it affects the size of the impedance loop being created. A good example of this is provided, precisely, in Fig. 4.2, wherein the LC pairs are distinct for each matching circuit as depicted in Fig. 4.1. The use of a shunted capacitor and inductor, considering the Smith chart, could seem pointless as the impedance follows circles of constant admittance, simply adding the inductor makes the impedance go upwards, and adding the capacitor makes the impedance go downwards, both going tangent to the mentioned circles. Therefore, it would appear that no significant alteration occurs—other than the apparent improvement of the impedance matching due to power being trapped in the matching network itself because of two additional lossy components—however, as lower and higher frequencies are affected more or less by capacitances and inductances, the curve of the impedance may close itself if the components are chosen wisely.

The LC resonator may be as lossy as the series inductor, which is the component generating the major power dissipation. Generally, the closer the impedance loop is made, the more lossy the LC resonator. The series capacitor practically does not dissipate power, and its usage is reserved for the cases in which the impedance loop is not well centered on the Smith chart, for which case an added capacitance may tune the position of the loop and further improve the impedance matching.

The losses incurred by the series inductor and LC resonator may be observed in Figs. 4.6 and 4.7, where the losses of the brown-colored (triangle) and green-colored (circle) matching networks from Fig. 4.1 are broken down to the two main parts of the broadband matching circuit. For the $L_s$ and $L_pC_p$ cases according to the legend, all the components of the circuit but the one indicated are lossless, and the indicated one is lossy; the remaining two curves indicate the simulated $S_{21}$ parameter for a completely lossless or lossy circuit.

Other than simplifications such as considering that the Q factor for each component is the same within the whole bandwidth—although the changes in the 824-960MHz band are on the order of 2% at most,—this representation is not entirely accurate as when one component is lossy and the others are not, the impedance matching is altered. Since the impedance when the components are lossless or lossy is different, the impedance matching is different as well. Nevertheless, it is estimated that this does not affect the simulations significantly and that the results are close to what would really happen.

Finally, in the high frequency region, two and three component topologies usually are enough for providing a good impedance matching with high antenna efficiencies. The two-component circuits are L topology ones (Fig. 4.8), and the three-component circuits compose a T topology (Fig. 4.9).

No particular component from any of these two topologies affects more than the others from a power dissipation point of view as the values necessary to provide good impedance matching tend to feature very similar Q and ESR parameters.
4.4. Matching Networks for Ground Plane Boosters

Figure 4.6: Rectangular chart showing the simulated $S_{21}$ parameter of a radiating system comprising a cube-shaped booster sized 5x5x5mm$^3$ using brown-coded matching network from Fig. 4.1 for the low frequency region. Averages of $S_{21}$ parameter in linear: lossless circuit 87%, lossy circuit 78%, lossy series inductor 80%, lossy LC resonator 85%.

Figure 4.7: Rectangular chart showing the simulated $S_{21}$ parameter of a radiating system comprising a cube-shaped booster sized 5x5x5mm$^3$ using green-coded matching network from Fig. 4.1 for the low frequency region. Averages of $S_{21}$ parameter in linear: lossless circuit 88%, lossy circuit 75%, lossy series inductor 81%, lossy LC resonator 82%.
4.5. Conclusions

A brief review of the two approaches for designing matching networks has been presented together with schematics for the matching network topologies that are generally used with ground plane boosters.

From the results of the simulations, it is apparent that the transducer power gain based designs provide a superior performance in comparison to the mismatch loss based ones.

The negative side of TPG focused networks is that they are not easily synthesized: the equipment for measuring the performance of the radiating system including a matching circuit does not provide direct information related to power dissipation and how to improve further the performance of the radiating system. In principle this problem also exists for mismatch loss based matching circuits because the information given by the equipment is the same, yet from an $S_{11}$ point of view, the engineer can infer which reactive components add or modify so as to keep minimizing the reflection coefficient.

So, in the end, the two approaches are usually combined: first, a matching circuit is made with the target of minimizing the reflection coefficient and, afterwards, adjustments on the components are made so as to seek the ones that make a less lossy system with the gain achieved not being lost due to a reduction in $S_{11}$. The second part may be done more easily taking into consideration the $Q$ factors specified by the lumped components manufacturer and the antenna efficiency measurements resulting from previous changes in the matching circuit.
Chapter 5

Ground Plane Boosters with a Bar Form Factor

5.1. Introduction

Cube-shaped ground plane boosters have been already developed in the latest years [1][19][20][25]. A particular radiating system employing two of such boosters, each one sized 5x5x5mm$^3$, and a ground plane with a 120x60mm$^2$ size, is able to provide at least dual-band operation in a low frequency region (LFR), in which the GSM850 and GSM900 cellular standards are typically comprised, and penta-band operation in a high frequency region (HFR), in which at least CDMA1700, GSM1800, GSM1900, WCDMA2100 and LTE2600 are within. Further radiating systems might use two additional and identical boosters put together as a single booster for operating the LTE700 cellular standard. The volumes it takes to put said boosters are, in total, 250mm$^3$ in the first example and 500mm$^3$ in the second example.

Albeit the volume used is quite small, there is an interest in boosters that may provide similar performance throughout the frequency bands with a reduced dimension in one of their axes: the height. One of the devices in which radiation boosters could potentially perform well are cellular phones; in spite of the trend of increasing the size of the phone in the so-called smartphone era, the dimensions that have seen a growth are the width and the height. In contrast, the thickness of the phone is being reduced so as to make it as slim as possible. Even though the height of the device increases, the height of the boosters in this thesis refers to the dimension that goes along the thickness of the device; the height of the booster is seen as the dimension that goes out of the ground plane along its normal. The notation format for booster dimensions used throughout the thesis is $WxHxL$, so the first dimension corresponds to the width of the booster, the second one corresponds its height, and the third one to its length.
Shrinking the height of the booster is thought to come at the expense of an increase in the other two dimensions. It has been stated that ground plane boosters mainly couple energy to or from the ground plane and the task of radiation is solely left to the ground plane, nevertheless the shape and size of the booster will affect the modes excited and how well they are excited on the PCB, so changing the form factor is not trivial.

This chapter is structured in two main sections. First of all, the effect of changing each dimension is analyzed and compared to the bandwidth of the cubic ground plane booster. Once a trade-off between dimensions that performs well in all three frequency ranges individually (single-port) has been set, performance is checked on a real prototype. The simulations are translated to an FR4 printed circuit board and the S parameters, antenna efficiency and radiation patterns are then measured, which will define how well the booster behaves. During the prototyping stage, several matching networks are tested according to the knowledge acquired in the previous chapter as well.

At the end, the final prototype is presented together with the measured parameters and installed matching networks.

5.2. Study of Form Factors

As it would seem prior to any simulation, a reduction in height has to imply an increment in width or length. Throughout this chapter, the dimension being parallel to the longer edge of the ground plane is referred to as the width of the booster, whereas the one parallel to its shorter edge is referred to as its length. The reduction in height is clear: one third of the original one is imposed to be gone.

To study the effect of the changes applied to the boosters shape and size, some parameter has to be computed or extracted from the simulations so as to make the comparison with the cubic booster possible, and guide the subsequent modifications until some dimensions seem to provide a good performance. The parameter chosen for this analysis is the inherent bandwidth defined in section 2.3, particularly in equation (2.5).

5.2.1. Starting Point for Design and Reference Values

The procedure for finding out an appropriate form factor is based on IE3D® simulations. As aforementioned, the new ground plane booster is expected to perform similarly to the cubic booster. So, for comparison purposes, a model of a booster placed in a ground plane corners is first built in IE3D®; said booster has 5x5x5mm³ dimensions, the ground plane sized has 120x60mm² dimensions, and the substrate layer defined is 1mm thick with the characteristic values that correspond to FR4 (the material which will be used for prototyping), namely a dielectric constant $\varepsilon_r=4.15$ and a loss tangent $\tan\delta=0.013$. The
substrate is, of course, replicated throughout the whole study.

The ground plane and a pad for allocating the booster are then included in the modeled substrate. The size of the pad corresponds to the footprint of the booster plus a 2x1mm² pad extension which is used for soldering the first component of the matching network. The separation between the ground plane and the entire pad is 0.5mm, interpreted as the shortest distance between one ground plane edge and the pad edge closest to the former. The positioning and separation of the pad are key in the excitation of the ground plane radiation modes.

Within the context of developing a new form factor, it is convenient to define a depth parameter that quantifies the distance between the edge of the ground plane and the farthest away edge of the booster pad. This parameter is a helpful indicator of the free space outside the ground plane necessary to allocate the booster. The depth of the cubic booster is 6.5mm because of the following elements between the edge of the ground plane and the distant-most part of the pad: the separation (0.5mm), the pad extension for component soldering (1mm) and the remaining part of the pad where the width of the booster has to fit in (5mm). Even though additional pads for soldering components are included when PCBs are actually made, they are included inwards the ground plane and thus the depth parameter remains unchanged. Regarding this, for the computation of the inherent bandwidth $BW_Z$, the model used herein does not include the rest of the pads for allocating the matching network because they do not change significantly the result. Moreover, the presence of such pads would have unnecessarily increased the number of unknown values in the _IE3D_ simulation, resulting in a longer computation time.

Besides, an additional parameter is defined: the gap of the booster. This parameter quantifies the distance between the edge of the ground plane and the side of the booster closest to the ground plane; this is equivalent to sum the separation and the pad extension, so the reference case has a gap of 1.5mm. The definition of these dimensions may be observed in Fig. 5.1.

![Figure 5.1: Definition of booster dimensions](image)

The fundamental limits on small antennas establish a direct relationship between their
size and their limit achievable bandwidth. The smaller the antenna is in terms of the wavelength, the narrower the achievable bandwidth becomes. Consequently, the most severe bandwidth conditions in this thesis apply to the longer wavelengths or, in other words, the lower frequencies. Although the ground plane acts as the main radiator, it also features stringent constraints in said lower frequencies. Therefore, the form factor study focuses on the low frequency region first. Since studying one booster instead of two is simpler, the initial attempts are bent on getting a form factor suitable for the GSM850/900 band and, later, find out which booster in LTE700 has an inherent bandwidth similar to the reference one that results from allocating two cubic boosters together. In between, the high frequency region is assessed in terms of bandwidth as well.

Once this is all set, the input impedance computed at the port is retrieved from \textit{IE3D}™ and directly input to equation (2.4) the output of which is the quality factor of antenna, and by means of equation (2.5), setting SWR or S to be 3 (the minimum necessary level to consider impedance matching and, thus, low-level signal reflections,) the inherent antenna bandwidth is computed. Nevertheless, the quality factor of the antenna \( Q_Z \) is a function of the frequency due to the change in impedance as the frequency varies. And since the bandwidth relates to \( Q \) (in this case \( Q_Z \)), the bandwidth is also dependent on the frequency. For simpler quantification and straightforward comparison, the inherent bandwidth is calculated just for one single discrete value of frequency: the central frequency. It has been considered that the variation in percentage which may happen at the central frequency will be, more or less, similar at the rest of frequencies of the band. In addition, it has to be noted that (2.4) includes derivatives of the impedance of the antenna—radiating structure in this case,—thus the impedance cannot be retrieved for the central frequency only but for a wide enough margin of frequencies. So the following has been done: the simulated input impedance for the whole bandwidth has been collected, with a step size smaller than 5 % of the entire frequency bandwidth; the function for the entire bandwidth has been evaluated and, finally, the inherent bandwidth just for the central frequency has been kept.

Having modeled and simulated the cubic booster in a PCB as described above, the reference inherent bandwidth that will be used across the form factor study is 13.2 % for the low frequency region delimited by the lower and upper frequencies: 824MHz and 960MHz; so the 13.2 % value corresponds to the inherent bandwidth at 892MHz.

At this point it is possible to start redefining the dimensions of the booster, and just reevaluating the quality factor of the structure with the input impedance of the reshaped booster. Computing the corresponding inherent bandwidth is enough to compare its potential performance with the potential performance of the \( 5 \times 5 \times 5 \text{mm}^3 \) booster.

Due to design requirements, a height of a booster reduced by one third has been established, so the new height is 3.2mm. Therefore, the starting point in the form factor study consists in modeling a replica of the \( 5 \times 5 \times 5 \text{mm}^3 \) ground plane booster, but changing its
height to 3.2mm. From there, changing iteratively the size of the booster in the remaining two dimensions and evaluating the inherent bandwidth will follow. There is an additional limitation though as the depth dimension has to be kept to the same value as for the reference booster, thus not requiring additional free space beyond the ground plane.

Testing a replica of the cubic booster just shrunk in height results in an inherent bandwidth of 10.7%. This means that, if a percent in reduction was to be computed, the booster would have its inherent bandwidth decreased by a 19% even though its volume is 36% smaller: from 125mm$^3$ down to 80mm$^3$.

5.2.2. Changes in Length and Width Dimensions of Booster

Given the requirement of a specific height and the constrain in depth, the two main dimensions which may be played with are the length and the width of the booster. Clearly, if there is a reduction in inherent bandwidth as the height gets reduced, and considering that the depth should not exceed the original one, the width can only be decreased which in principle would mean losing further bandwidth.

So, apparently, the best dimension to start the modification with is the length of the booster. If boosters that have only experienced discrete increments in length are simulated, it is possible to plot the trend of the inherent bandwidth as the length is swept as in Fig. 5.2. A first batch of simulations is run with three different booster sizes: 5x2.5mm$^2$, 5x3.75mm$^2$, and 5x5mm$^2$; all of them having the length as the remaining dimension. The first size corresponds to the cubic booster with half the height and prior to the length variation, the last size is the original reference booster, and the third booster size is the booster with height just in-between the two. The remaining dimension is what the inherent bandwidth will depend on as it is made to be a function of said length. The simulations were performed in the low frequency region, and the inherent bandwidth (2.5) calculations are represented for 892MHz.

As knowledge regarding the effect that each dimension has on $BW_Z$ is sought, the boosters being studied do not comprise any size with a 3.2mm-height as that is to be introduced later when the form factor is actually designed.

As expected, enlarging the length results in an increase in bandwidth. Moreover, it can be appreciated that there is a linear relationship between the length of the booster and the inherent bandwidth, and that the slope of the curve is not exactly identical between one initial size and the other. Albeit the volume changes in the same proportion when considering each booster independently, the increase in volume in one booster is the double of the other one but the slope is not. Therefore computing the effect of varying the length for one booster does not allow extrapolating which effect the same variation would have for another booster unless it is very close in dimensions.
This first simulation supports the hypothesis of longer boosters certainly improving the inherent bandwidth parameter, which helps in the design process in case that a specific figure in terms of bandwidth had to be reached.

Additionally, judging by the curves and making the alternative reading, the reduction in width has a larger negative impact the bigger the booster is: for a 5mm length, the difference in bandwidth is around 4.5%, whereas for a 10mm length, the difference is over 5.5%.

After length, width is the dimensions to be swept next. The limits of the dimension are directly chosen from the entire and half cubic booster, so 5mm and 2.5mm respectively. And three curves are again plotted representing the inherent bandwidths for boosters sized 5x5mm$^2$, 5x6.25mm$^2$, and 5x7.5mm$^2$; with the remaining dimension being the width that is swept. The results from the simulation are shown in Fig. 5.3.

Once again, the relationship is linear and decremental: as the width decreases so does the bandwidth. Although it may be difficult to notice, the curves are not completely parallel, something that was expected as the inherent bandwidth plot versus length plot already pointed in that direction: the variation in width results in non-constant variations of bandwidth for different booster sizes.

The graph for now does not go further than 5mm because of the depth restriction. It is known that the gap dimension must be, at least, 1.5mm for the allocation of the first component of the matching network and, thus, it is not possible to increase the width keeping the same depth.
5.2. Study of Form Factors

5.2.3. Changes in Gap Dimension of Booster

It has been shown what is the effect of altering either the length or the width of a booster is. It would seem rather obvious to compensate the reduction in bandwidth, caused by the loss of height, increasing the length of the booster because any reduction in width would further decrease the bandwidth. There is, however, a further dimension which may be modified: the gap.

As the booster is in charge of the excitation of ground plane modes, it is apparent that its form and size matter. But so does its location. As documented, a capacitive radiation booster performs better at the corner of the ground plane, the corner being in the vicinity of a shorter edge of the ground plane. In fact, the whole study of the new form factor is based on the best possible location for a capacitive booster, still how the gap affects the excitation of modes and, subsequently, the inherent bandwidth, remains unknown.

The minimum gap must necessarily be 1.5mm, while the maximum gap is limited by the depth and the width of the booster. If the depth is fixed, the only possible option is to reduce the width and lose bandwidth in order to increase the gap. For the sake of knowing the relevance of the gap, the width of the booster is again reduced to compensate for the increase in gap.

In Fig. 5.4, distinct curves corresponding to fixed values of width represent the inherent bandwidth of a booster as its gap is swept, hence the depth dimension is always 6.5mm or less.

As it may be appreciated, the larger the gap is for any of the simulated boosters, the
5.2.4. Optimal Pair of Gap and Width Dimensions

So far the relevance of up to three distinct dimensions has been reviewed: the length, the width and the gap. With regard to the width and the gap, the data from the simulations corroborated that the inherent bandwidth (most likely) improves when any of such dimensions is enlarged.

Judging by the results obtained in section 5.2.3, the fact there is an optimal pair of gap and width values for a booster given a specific depth can be concluded. Going back to Fig. 5.4, differences in the resulting bandwidth for the same depth can be noticed. The goal now is to find out which combination of gap and width maximizes $BW_Z$ among all boosters.

By means of new batches of simulations where gaps and widths are swept simultaneously while maintaining a constant depth at all times, the optimal gap-width pair may be found (see Fig. 5.5). The booster simulated this time features width and length of 5mm each.

It is clear that there are maxima within the inherent bandwidth curves represented in function of the gap and width of the booster. For different depth values, then, a different
5.2. Study of Form Factors

Figure 5.5: Simulated inherent bandwidth (for SWR ≤ 3, at f₀=892MHz) as the gap and width for a 5mm-wide and 5mm-long booster are varied, for depth dimensions equal to 5.0mm, 5.5mm, 6.0mm, and 6.5mm

pair achieves the best result. It also demonstrates that maximizing the size of the booster does not achieve the best results, which further shows that ground plane boosters indeed excite the ground plane.

5.2.5. Candidate Form Factors for Booster

Knowing there is a combination of gap and width which, in the current form factor study, is expected to achieve better results for the same depth compared to other combinations, such a value combination is first chosen to establish the aspect ratio for a new ground plane booster.

If the depth is kept to 6.5mm as in the original PCB comprising the 5x5x5mm³ boosters, the optimal gap and width have roughly the same value: about 3.25mm each. This means that the booster would be as far from the ground plane as its own width or, in other words, a second booster would perfectly fit between the edge of the ground plane and the booster spaced 3.25mm from said edge.

Considering these values, it is strictly possible to use them for a new booster solution: a 3.25mm-wide booster manufactured and placed on a PCB. However, from a technical point of view, that would not be a wise idea for two main reasons. First, current manufacturing techniques are neither accurate nor precise enough for cutting boosters up to hundredths of a millimeter; in fact, the tolerances in this kind of processes are between one and two tenths of a millimeter, otherwise the costs would increase substantially. And secondly,
setting booster dimensions other than its height to values also similar to 3.2mm could lead to wrong-sided placements on the PCB. Even though said wrong placements could be avoided by marking one or several sides with ink, it is more convenient to set the width of the booster to 3.2mm like its height.

A first form factor, then, would be 3.2x3.2x5mm$^3$, wherein the length is the same as the cubic booster. As the 6.5mm depth is kept, the gap is 3.3mm. The inherent bandwidth for this new set of dimensions, for the central frequency of the 824-960MHz band, is now 10.9%, quite far from the 13.2% achieved by the original radiation booster. Interestingly, as it was computed before, the inherent bandwidth of a booster having dimensions 5x3.2x5mm$^3$ is 10.7%, which is less than the value achieved by a smaller booster occupying the same footprint on the PCB.

Based on the simulations performed in section 5.2.2, further simulations have to be run for the proposed booster changing its length. Having two distinct length values with their corresponding bandwidths is enough to infer the length necessary for reaching the 13.2% or for at least being very close.

With an inherent bandwidth of 13.1%, the required length according to the extrapolation would be 8mm— with the result being confirmed with a dedicated simulation with said length. Even though said length would be necessary to have similar $BW_Z$ with respect to the 5x5x5mm$^3$ booster, a first proposed form factor for testing purposes is 3.2x3.2x7mm$^3$, these dimensions resembling to the shape of a bar. These dimensions, which are characterized by a volume of 72mm$^3$, imply falling short in terms of $BW_Z$ (12.5% against 13.2% of the original case,) yet the inherent bandwidth has not been calculated for the LTE700 band and the high frequency region, so how it compares to the cubic booster in these two frequency ranges remains unknown.

A 120x60mm$^2$ ground plane is now modeled in **IE3D**—with two 5x5x5mm$^3$ boosters put together having the same characteristics the single booster model had, that is a 0.5mm separation, a 1mm pad extension, and a 5mm wide booster pad, the pad now having a length of 10mm. This configuration is only used for operation in the LTE700 band, which ranges from 698MHz to 798MHz. For the high frequency region delimited by 1.71GHz and 2.69GHz, the initial model is enough as a single booster suffices for providing a wide enough bandwidth that covers said range of frequencies. Computing $BW_Z$ for the central frequencies once again, the original booster achieves 7.0% for the LTE700 band (the central frequency being 748MHz) in a two-booster configuration, whereas the resulting $BW_Z$ for the high frequency region (at 2.2GHz) is 42.5%.

Computing the same parameter now for the 3.2x3.2x7mm$^3$ booster, the values are 7.2% and 39.0% for LTE700 and the high frequency region, respectively. So, summarizing, this first candidate features a slightly greater inherent bandwidth in the LTE700 band; and, in contrast, a slightly lower one in the low and high frequency regions. Based on the
5.3. Simulated Performance of Booster Candidates

The performance of the cubic booster in the 1.71-2.69GHz band, the lower $BW_Z$ of this new form factor is not expected to present problems in terms of impedance bandwidth; however the $S_{11}$ parameter in the low frequency region is expected to be greater than desired.

In addition, as the volume has been reduced a 42% with respect to the cubic booster, a second form factor is proposed as a back-up: 3.2x3.2x10mm$^3$—also resembling the shape of a bar—for a volume of 102mm$^3$, an 18% less than the original 125mm$^3$. It would be possible to reduce length further down to 8mm, the first one to reach the inherent bandwidth in the low frequency region, or even 9mm, but an even better antenna efficiency is expected from a 10mm-long booster which still has a volume smaller than that of the cubic booster. Therefore, both candidates (7mm-long and 10mm-long) are to be tested before deciding which one will be proposed as the final booster with a bar form factor.

5.3. Simulated Performance of Booster Candidates

At this point, two bar-shaped candidates have been presented and are to be tested via further simulations. Previously, the simulations were performed focusing on the calculation of the $BW_Z$ parameter. In the current section, the simulations shall include matching networks with ideal and lossless components as a first estimation of the boosters performance in terms of antenna bandwidth. The models generated for the inherent bandwidth computation are reused in these simulations as one-port black boxes connected to the matching circuit.

The following subsections show the impedances of the boosters with matching circuits specifically synthesized to minimize the reflection coefficient within the particular frequency regions of interest, in pursuit of maximizing the power transferred to the load or, in other words, maximizing the antenna efficiency. As the components are said to be ideal and lossless, only mismatch loss based designs are synthesized. The matching circuits comprise, at most, four components in topologies which may be installed in a PCB including two component-mounting pads per booster, where the matching networks have been presented in section 4.4. In the successive subsections, the same structure is kept: first the cubic booster is presented and, afterwards, the 7mm-long and 10mm-long boosters are presented as well, so their performance might be compared to the original booster. Finally, a single rectangular plot representing the $S_{11}$ for the three boosters is shown for straightforward reflection coefficient comparisons.

Please note that the schematics for the matching circuits tested in this section are included in Appendix A. The used circuits are indicated in the caption of each figure.
5.3.1. Low Frequency Region

The first band to be studied is the 824-960MHz frequency range which generally includes the GSM850 and GSM900 cellular standards.

The circuit being implemented is a broadband matching network with the addition of a series capacitor at the feeding terminal which will help at the prototyping stages for finely adjusting the impedance and, thus, minimize the reflection coefficient. So, for reference purposes, the impedance at the feed port for the cubic booster using a broadband matching circuit is shown in Fig. 5.6.

![Smith chart showing the simulated input impedance, in the low frequency region, of the reference cube-shaped booster sized 5x5x5mm^3 using the matching network from Fig. A.1 considering ideal lossless components](image)

Figure 5.6: Smith chart showing the simulated input impedance, in the low frequency region, of the reference cube-shaped booster sized 5x5x5mm^3 using the matching network from Fig. A.1 considering ideal lossless components.

Now, the attention is focused on the booster candidates. First, the booster candidate with 3.2x3.2x7mm^3 dimensions features the impedance from Fig. 5.7 using a different broadband matching network. The $S_{11}$ parameter is between -8dB and -10dB, a quite good result in spite of a 0.7% negative difference in inherent bandwidth with respect to the 5x5x5mm^3 booster. The bandwidth is correct for operating as per cellular standards, still pending to verify its antenna efficiency with the matching network in place.

The 10mm-long case, on the contrary, achieves a sensitively better match (Fig. 5.8) as predicted beforehand. There is an interesting difference with respect to the other booster candidate: the series inductors used by each one of them. The former candidate uses a series inductor of 51nH (see the matching circuit from Fig. A.2) for partially cancelling the capacitive behavior of the booster, whereas the latter uses a significantly smaller series
5.3. Simulated Performance of Booster Candidates

Figure 5.7: Smith chart showing the simulated input impedance, in the low frequency region, of a bar-shaped booster sized 3.2x3.2x7mm$^3$, with a 3.3mm gap, using the matching network from Fig. A.2 considering ideal lossless components.

The difference in series inductors is particularly relevant because the Q parameter for is over 10% worse for the 51nH with respect to the 43nH (considering high-Q 0603 SMD components.) Even if the Q were to be the same for both inductors, the dissipated power would already be quite larger for the bigger inductor because of a larger ESR involved. This fact, in a real case, would translate into a significantly improved match for the 7mm-long candidate due to the losses of the component. Albeit prior to the simulation the fact that the longer booster would have a superior performance was predictable to a certain point, this superiority is two-fold: firstly, the $S_{11}$ is lower in an ideal lossless scenario and, secondly, once the losses are taken into account, less power is lost in terms of heat, fortunately meaning that a larger percent of the power is transferred to the load.

As an additional remark, there is an astonishing similarity between the impedance of the cubic booster and the 10mm-long candidate as seen in Figs. 5.6 and 5.8. This can also be appreciated in the rectangular graph from Fig. 5.9 which compares the $S_{11}$ parameters of the candidate boosters and the reference one.

Nevertheless, a closer look to the matching networks reveals a likely sub-optimal antenna efficiency due to very lossy components. A TPG design as described in chapter 4 shall
Figure 5.8: Smith chart showing the simulated input impedance, in the low frequency region, of a bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. A.3 considering ideal lossless components.

Figure 5.9: Rectangular chart showing the $S_{11}$ parameter, in the low frequency region, of the two bar-shaped booster candidates and the reference cube-shaped booster with corresponding matching networks considering ideal lossless components.
be considered instead. Moreover, the example from Figs. 4.2 and 4.3 have shown how not so closed loops may perform better. For now, the results presented in this section are kept as they are as both the booster candidates and the reference booster have been tested in the same conditions.

5.3.2. High Frequency Region

The high frequency region packs a really large bandwidth with almost 1GHz for comprising several cellular standards. Although one could also find the frequencies for WLAN, commonly referred to as WiFi connectivity (2.4-2.5GHz), in practice an antenna for cellular services is not used for WLAN, rather a separate antenna is provided. Recently dedicated WiFi antennas have been made dual-band so as to provide operability in the 5GHz spectrum (IEEE 802.11a standard) as well. Regarding cellular connectivity, the main standards are allocated in the lowest part of this region, namely 1.71-2.17GHz; these standards are mainly 3G although some countries are also reusing them for LTE or 4G services.

As it occurred in the low frequency region, the 3.2×3.2×7mm³ booster is characterized by a slightly lower $BW_z$ than the 5×5×5mm³ booster, although it manages to get an $S_{11}$ below -8dB anyway with an impedance loop well positioned in the center of the Smith chart as seen in Fig. 5.11, whereas the cubic booster presents the response in impedance depicted in Fig. 5.10.

![Smith chart showing the simulated input impedance, in the high frequency region, of the reference cube-shaped booster sized 5x5x5mm³ using the matching network from Fig. A.4 considering ideal lossless components](image)
The 10mm-long booster, on the other hand, achieves an $S_{11}$ in the order of -10dB within the whole band. Contrarily to the low frequency region, no apparent change in terms of power dissipation is expected to occur in this band judging by the values initially synthesized for the circuit.

The performance of either candidate is correct in terms of reflection coefficient. Albeit based on the Smith charts the 10mm-long booster seems to have a better response than the original cube-shaped booster, the representation of all reflection coefficients in a rectangular diagram shows rather similar curves between all boosters (Fig. 5.13). An important difference though is the up to 1.5dB $S_{11}$ variation between the two proposed form factors at the lowest frequencies of the band. It is convenient to remember that the most-used cellular standards in the high frequency region are located in that part of the band. This, in the end, may be a key point for deciding which booster size shall be used in the prototyping stage.

### 5.3.3. LTE700 Band

Finally, the LTE700 band going from 698MHz to 798MHz is studied, the name of the band being directly gathered from the 4G cellular standard that is operated. At these frequencies it is necessary to put two boosters together side by side for an improved performance, which translates into enough antenna bandwidth for giving operation to the
Figure 5.12: Smith chart showing the simulated input impedance, in the high frequency region, of a bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. A.6 considering ideal lossless components.

Figure 5.13: Rectangular chart showing the $S_{11}$ parameter, in the low frequency region, of the two bar-shaped booster candidates and the reference cube-shaped booster with corresponding matching networks considering ideal lossless components.
Chapter 5. Ground Plane Boosters with a Bar Form Factor

LTE standard and theoretically enough antenna efficiency for transmission and reception of signals.

The matching circuit being used is a series inductor plus a pi-network, namely a 4-component ladder circuit. In the LTE700 band it is necessary to first add an inductance cancelling the highly capacitive component of the booster in the operating frequencies, as it already occurred in the low frequency region. Special care has to be taken in this regard because the series inductor performing such task may be the main loss-introducing component in the matching network overall.

Placing the two cube-shaped boosters and an $S_{11}$-optimized matching network, the simulated impedance seen at the feeding port is represented in Fig. 5.14.

![Figure 5.14: Smith chart showing the simulated input impedance, in the LTE700 band, of two reference cube-shaped boosters sized 5x5x5mm³ put together and using the matching network from Fig. A.7 considering ideal lossless components](image)

The setup including two 7mm-long boosters connected to an ideal ladder circuit is characterized by the impedance shown in Fig. 5.15.

The worst $S_{11}$ achieved in the band is -6.7dB. A quite open impedance loop can be observed and, although it is does not overlap the SWR=3 circle, it is indeed quite close to it. Still, the response achieved by this pair of boosters is better than that of the two side by side cubic boosters, which had a worst $S_{11}$ of -5.8dB and an equally open impedance loop.

Putting two 10mm-long boosters together now and matching them in impedance with a ladder network, the impedance shown in Fig. 5.16 is measured at the feeding port. The
5.3. Simulated Performance of Booster Candidates

Figure 5.15: Smith chart showing the simulated input impedance, in the LTE700 band, of two bar-shaped boosters sized 3.2x3.2x7mm$^3$ put together, with a 3.3mm gap, and using the matching network from Fig. A.8 considering ideal lossless components. The impedance loop in this case is significantly smaller, and the worst $S_{11}$ within the whole band is -8.3dB.

Figure 5.16: Smith chart showing the simulated input impedance, in the LTE700 band, of two bar-shaped boosters sized 3.2x3.2x10mm$^3$ put together, with a 3.3mm gap, and using the matching network from Fig. A.9 considering ideal lossless components.
Comparing the two matching circuits (Fig. A.8 and Fig. A.9), there are no big differences other than the series inductor—the longer booster features a slightly less capacitive input impedance and, thus, a 36nH inductor suffices for canceling the imaginary part of the impedance, whereas the shorter booster requires a 39nH inductor for achieving a similar result. Obviously, there is an impact for using a larger inductance because of its typically lower Q factor, which affects the power lost in the network as already described. However, the difference in this case is not as big as it was in the low frequency region. Anyway, there is a relevant difference in terms of $S_{11}$ between the two dimensions as shown in Fig. 5.17, which is to be considered for deciding the final candidate for the prototyping stage.

![Figure 5.17: Rectangular chart showing the $S_{11}$ parameter, in the LTE700 band, of two of either bar-shaped booster candidates and reference cube-shaped boosters put together with corresponding matching networks considering ideal lossless components](image)

### 5.3.4. Conclusions

After having simulated the different frequency regions in which the newly-shaped boosters have to operate, it is convenient to summarize the results they achieved in order to decide whether any of them will prevail to be tested in real prototypes.

The proposed candidates had dimensions of $3.2\times3.2\times7\text{mm}^3$ and $3.2\times3.2\times10\text{mm}^3$. The larger booster proved to have more potential in terms of inherent bandwidth in each of the three frequency regions. This, in turn, also paid off in terms of reflection coefficient and reduced expected losses due to the $Q$ of the lumped components used in the matching networks.

The most significant difference is present in the low frequency region, wherein the $S_{11}$
and the series inductor seem to anticipate a relatively better antenna efficiency for the 10mm-long booster than for the 7mm-long booster. In addition, the $S_{11}$ of the 7mm-long booster compared to the cubic booster was slightly worse.

Besides, the 10mm candidate provides a better bandwidth than both 5x5x5mm$^3$ and 3.2x3.2x7mm$^3$ boosters in the LTE700 band in a two side-by-side configuration, the range of frequencies in which it is very difficult to get good response in terms of reflection coefficient and antenna efficiency.

Considering these two aspects, it has been decided to use, in the prototyping stage, the booster with dimensions 3.2x3.2x10mm$^3$. The volume of this booster is slightly smaller than the cubic one—102mm$^3$ versus 125mm$^3$—so a similar or even better performance is expected with a lighter ground plane booster featuring two thirds of the original height.

5.4. Measured Performance of Booster Candidate

5.4.1. Introduction

The performance of the two booster candidates was proven successful based on the simulation results. Although the capacitors and inductors in the simulations were lossless, other than already described power dissipation characterized by quality factor $Q$ and equivalent series resistance ESR of lumped component, many more undesired phenomena have been neglected. Among these phenomena are the tolerance of the components, parasitic inductances or capacitances leading to component resonances, nor different effective reactances at each frequency, the effect of the pads for soldering lumped components. All these are to be taken into account and measured in terms of $S_{11}$ and antenna efficiency now that real prototypes are being built.

So, even though the two booster candidates seemed to perform fairly well, the best-performing candidate is to be prototyped to compensate for the non-consideration of the aforementioned phenomena that simplified the computational burden. The performance of the booster this way is expected to be considerably satisfactory either because of meeting the minimum requirements at the limit frequencies and within the bandwidth, or because of having a better performance overall than the other candidate.

This decision does not necessarily mean that the booster featuring a length of 7mm would not provide a performance good enough for fulfilling the requirements imposed by the cellular industry standards. A solution that more predictably works according to the criteria established throughout this thesis is rather sought.

Therefore, the following subsections present the performance of brass-made boosters sized 3.2x3.2x10mm$^3$, such as the ones shown in Fig. 5.18, with a 3.3mm gap using an FR4 printed circuit board in which the ground plane has a size of 120x60mm$^2$—the size of a
common smartphone.

Figure 5.18: Photo of brass-made boosters sized 3.2x3.2x10mm$^3$ used in the prototyping stage.

The structure is as follows: first, single-port configurations are shown, thus only one booster is installed in the PCB at a time for testing its performance in a determined frequency region; for each particular frequency region, several matching networks are tested; afterwards, a triple-port prototype is presented using the final matching circuits.

Please note that every matching circuit used in any of the following sub-sections has its schematic drawn in Appendix B; each figure includes in its footnote the specific matching network reference and, for the antenna efficiency measurements, the respective average antenna efficiency measured within the bandwidth of interest.

5.4.2. Single-Port Configuration: Low Frequency Region

For the low frequency region (824-960MHz), it is deemed appropriate to use a broadband matching network which has been seen in the past as a reliable topology for a wide enough antenna bandwidth. For a -6dB $S_{11}$ threshold, it is possible to achieve a bandwidth wider than the necessary 15%, however it is more convenient to make the bandwidth narrower for getting a better reflection coefficient.

For these first prototypes, the printed circuit board is constructed with the PCB layout designed using AutoDesk® AutoCAD®. The layout is made so that it includes all the necessary booster pads for a triple-port configuration as seen in Fig. 5.19. This helps in saving time and substrate as the boosters and matching networks may be soldered and desoldered several times. Thus, it is possible to reuse the PCB just by making the required modifications at any time.

After a first prototype like the PCB from Fig. 5.20 is built, the input impedance at the feeding port of the radiating structure including a 10mm-long booster is measured (see Fig.
5.4. Measured Performance of Booster Candidate

Figure 5.19: *AutoDesk® AutoCAD®*’s 3D perspective of the triple-port PCB’s layout

5.21); for this measurement, the feeding port is maintained and the pads are short-circuited properly with 0Ω resistors and the parallel connection to ground are open-circuited.

Figure 5.20: Photo of a single-port PCB including three booster pads for bar-shaped boosters sized 3.2x3.2x10mm³; one ground plane booster is installed and connected to a port; the cable used is a microcoaxial and the connector is a 50Ω-SMA; the PCB is made of 1mm-thick FR4 with $\varepsilon_r=4.15$ and loss tangent $\tan\delta=0.013$; the ground plane has 120x60mm² dimensions.

The impedance at the feeding terminal is clearly capacitive with the resonance being located at 2.79GHz. This measurement reaffirms that an inductor is convenient at the beginning of the matching network for compensating the imaginary part of the impedance.
Therefore, using *IE3D* simulations is a good starting point, but it is also wise to use the direct measurement of the input impedance of the radiating structure for tailoring the matching circuit—with the necessary adjustments once the circuit is mounted so as to perform as expected—in *Microwave Office®*. All in all, it is possible to choose from several suitable matching circuits, so a few have been installed in printed circuit boards for measuring their response both in terms of reflection coefficient and antenna efficiency.

First, boosters with a matching network that uses a 39nH series inductor are built and measured. With this inductor, two distinct circuits have been tested, which mainly differ in the pair of values of their LC resonator, have been tested in an attempt to observe how small and big impedance loops affect the reflection coefficient and efficiency, wherein the smaller loop is represented in Figs. 5.22 and 5.23, and the bigger loop is represented in Figs. 5.24 and 5.25.

Albeit the reflection coefficient is not so good for the configuration with a bigger impedance loop, the antenna efficiency is three points superior (63% to 60%) to the better matched radiating system owing to a less lossy broadband resonator. These two structures clearly show why not only the impedance matching matters; typically, when two or more circuits present similar $S_{11}$ parameters, the one with more lossy components should be discarded, however this may also be the case when the $S_{11}$ is not that similar, as a less matched system still may transfer more power to the load.

These two examples perfectly depict a problem that was slightly appreciated in the
5.4. Measured Performance of Booster Candidate

Figure 5.22: Smith chart showing the measured input impedance, in the low frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.1

Figure 5.23: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the low frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.1. Average antenna efficiency $\eta_{a,avg} = 60\%$
Figure 5.24: Smith chart showing the measured input impedance, in the low frequency region, of the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, using the matching network from Fig. B.2

Figure 5.25: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the low frequency region, of the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, using the matching network from Fig. B.2. Average antenna efficiency $\eta_{a,avg} = 63\%$
5.4. Measured Performance of Booster Candidate

simultaneous stage, which is the fact that the edges of the low frequency region (824MHz and 960MHz) are not symmetrical in terms of reflection coefficient. This means that having a rather well centered impedance loop as in Figs. 5.22 and 5.24—the latter being slightly more shifted to the upper part of the Smith chart, yet moving it down would worsen the $S_{11}$ attached to the 824MHz,—the points delimiting the frequency range are not evenly distributed around the loop, not even close. It is not necessary that the two frequencies are in the curve intersection, but $S_{11}$-wise it would be convenient that they were equally as close or as far from said intersection. When this is not the case, one part of the bandwidth is better matched than the other resulting in a not so constant antenna efficiency, despite it would be desirable to have them as constant as possible—in cellular systems, narrow sub-bands assigned by the base stations are used by the handsets and, thus, it is not possible to give priority to a better performing sub-band as that solely depends on the base station.

In an attempt to move the upper and lower frequencies to a more suitable position on the impedance curve, the series inductor is now increased to 43nH. The added inductance shifts the resonance of the booster to a lower frequency and so do the frequencies from the band within the impedance curve.

Up to three distinct circuits have been built and measured, wherein the differences are basically the shunt capacitor-inductor pair that gives the form to the impedance loop. The fine-tuning capacitor also varies, although that is merely for centering the impedance on the Smith chart once the loop is defined.

The results are presented as follows: the smallest impedance loop performs as in Figs. 5.26 and 5.27, the response of a slightly bigger one is shown in Figs. 5.28 and 5.29, and the one with the biggest loop performs as shown in Figs. 5.30 and 5.31.

Again, it may be appreciated an uneven position of the 824MHz and 960MHz on the input impedance. In contrast to the circuits including a 39nH series inductor, which had the higher frequencies better matched, the ones with 43nH inductor provide a superior impedance matching to the lower part of the bandwidth. It may be concluded that an inductor with an inductance in-between 39nH and 43nH would lead to an improved response because of a more uniform bandwidth. Unfortunately, there are no commercial inductors between these two values, so it is a matter of giving priority to one of the two edges of the band and choose the component accordingly. Even if there was an inductor of about 41nH, it would be necessary to check its Q factor because a better impedance matching in expense of much larger power dissipation would not be convenient as previously demonstrated.

From the last three configurations, the middle-sized impedance loop setup (Figs. 5.28 and 5.29) slightly outperforms the rest, and would even probably had a better antenna efficiency if the LC resonator and the series capacitor component values were finely adjusted, as the loop is shifted from the center of the Smith chart.

Anyway, both the matching networks using either 39nH or 43nH inductors have proven
Figure 5.26: Smith chart showing the measured input impedance, in the low frequency region, of the bar-shaped booster sized $3.2 \times 3.2 \times 10 \text{mm}^3$, with a $3.3 \text{mm}$ gap, using the matching network from Fig. B.3.

Figure 5.27: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the low frequency region, of the bar-shaped booster sized $3.2 \times 3.2 \times 10 \text{mm}^3$, with a $3.3 \text{mm}$ gap, using the matching network from Fig. B.3. Average antenna efficiency $\eta_{a,\text{avg}} = 55\%$. 
5.4. Measured Performance of Booster Candidate

Figure 5.28: Smith chart showing the measured input impedance, in the low frequency region, of the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, using the matching network from Fig. B.4

Figure 5.29: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the low frequency region, of the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, using the matching network from Fig. B.4. Average antenna efficiency $\eta_{a,avg} = 58\%$
Figure 5.30: Smith chart showing the measured input impedance, in the low frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.5

Figure 5.31: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the low frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.5. Average antenna efficiency $\eta_{a,avg} = 54\%$
that minimizing the reflection coefficient is not optimal, yet it is to be noted that tailoring matching circuits is not straightforward. The $S_{11}$ parameter may be easily measured by means of a network analyzer as the one used for this thesis, with said measurement taking not more than a minute. On the other hand, measuring the antenna efficiency for a bandwidth as relatively narrow as the low frequency region, in which a 5MHz-step is about right in terms of accuracy, takes approximately five minutes using the *Satimo SG 32*. Moreover, measuring the antenna efficiency does not provide any insight into how to improve it or where the power goes, whereas measuring the reflection coefficient one may see the input impedance and deduce which components could be modified for providing a better impedance match. It is also a fact that both impedance matching and losses are included in the $S_{11}$ measurement; the experience of the engineer developing the network and the radiating system is what provides additional insight into the reflection coefficient for assessing which parts of the $S_{11}$ correspond to the load being matched and which part corresponds to power dissipation.

The bottom line is that, other than using RF software tools such as *Microwave Office®* which may help in optimizing the matching circuits, the usual process for designing a network that provides a close-to-minimum reflection coefficient and fine-tuning some of its components afterwards so as to improve the efficiency. This way, it is possible to extract the data from the network analyzer, which can be easily interpreted for further modifications of the network, instead of trying to do so with the measurements taken in the anechoic chamber. Nevertheless, this approach may lead to a sub-optimal radiating system as the network that maximizes the transducer power gain might never be mounted.

5.4.3. Single-Port Configuration: High Frequency Region

After studying the performance of a 10mm-long bar-shaped booster in the low frequency region, it is time to do the same in the high frequency region which goes from 1.71GHz up to 2.69GHz, almost 1GHz or 44.5% in fractional bandwidth.

Based on the simulations, the $S_{11}$ of a bar-shaped booster should be capable of staying below -8dB for the whole bandwidth using a 3-component matching network as seen in Fig. 5.13. If this could be accomplished in a real prototype too, it would be important to see the effect on the antenna efficiency, although low efficiency would be unexpected because the components are not particularly lossy according to their quality factors.

A good starting point for designing the matching network is, once again, the input impedance with a booster coupled to the feeding port via $0\Omega$ resistors; as the high frequency region only requires one booster, the initial measurement presented in the previous subsection is also valid (Fig. 5.21).

Starting with a T-topology matching network synthesized using *Microwave Office®*,
components are selected so that the input impedance is compressed and kept close to the center of the Smith chart for a better reflection coefficient. Accordingly, a particular tested network makes the radiating system to perform as in Figs. 5.32 and 5.33.

![Smith chart showing the measured input impedance, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, using the matching network from Fig. B.6.](image)

**Figure 5.32**: Smith chart showing the measured input impedance, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, using the matching network from Fig. B.6

An average antenna efficiency of 74% is quite high; however, as in the previous matching network the series inductor is only 1.5nH—slightly decreasing the resonant frequency,—an L-topology matching network obviating the series inductor could possibly achieve a fairly similar $S_{11}$ not incurring in the power losses of a third component and making the radiating system more robust against the tolerances of the components.

For this reason, up to three distinct combinations of values in an L-topology matching network are built and measured. The differences among the three are just minor value adjustments of the components themselves. The measurements are presented in Figs. 5.34 and 5.35 for the first matching circuit, in Figs. 5.36 and 5.37 for the second one, and in Figs. 5.38 and 5.39 for the last one.

Very similar curves are obtained with all three L-topology matching circuits. It is concluded that the series inductor may be avoided, in the radiating systems considered herein with particular ground plane dimensions and sizes, and still have an average antenna efficiency above 75%. Not only the system is more robust against component value deviations, but it is also more efficient in $\eta_{\text{a}}$ as the T-topology achieved 74% in average, although the response could be possibly slightly improved by making minor adjustments.
5.4. Measured Performance of Booster Candidate

Figure 5.33: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.6. Average antenna efficiency $\eta_{a,avg} = 74\%$

Figure 5.34: Smith chart showing the measured input impedance, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.7
Figure 5.35: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.7. Average antenna efficiency $\eta_{a,avg} = 77\%$

Figure 5.36: Smith chart showing the measured input impedance, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.8
5.4. Measured Performance of Booster Candidate

Figure 5.37: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the high frequency region, of the bar-shaped booster sized $3.2\times3.2\times10\text{mm}^3$, with a $3.3\text{mm}$ gap, using the matching network from Fig. B.8. Average antenna efficiency $\eta_{a,\text{avg}} = 76\%$

Figure 5.38: Smith chart showing the measured input impedance, in the high frequency region, of the bar-shaped booster sized $3.2\times3.2\times10\text{mm}^3$, with a $3.3\text{mm}$ gap, using the matching network from Fig. B.9
Figure 5.39: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.9. Average antenna efficiency $\eta_{a,\text{avg}} = 78\%$

Overall, the performance of the booster in the high frequency region is considered to be satisfactory. It remains to be seen if the three-port configuration features the same reflection coefficient and efficiency, which ultimately will judge if the chosen bar-shaped booster operates in a satisfactory manner.

5.4.4. Single-Port Configuration: LTE700 Band

The last single-port configuration corresponds to the operation in the LTE700 band whose limiting frequencies are 698MHz and 798MHz.

As two boosters are put together in a single booster pad, the input impedance differs from the one previously shown. Placing the bar-shaped boosters in a prototype PCB and feeding them directly through $0\Omega$ resistors, the input impedance is measured and represented in Fig. 5.40 (only the impedance corresponding to the range of frequencies of interest is drawn.)

The impedance of the booster is largely capacitive, although having two boosters makes the impedance at the frequencies of LTE700 not as capacitive as if just one booster is used. Compared to Fig. 5.21, a single bar-shaped booster presents a reactance of about $-210\Omega$ at 960MHz and $-250\Omega$ at 824MHz. It is clear that the reactance in the 700-800MHz frequency range is even lower than that, and this would result in a serious penalty in terms of efficiency once a matching network trying to compensate for such capacitance is added.
Figure 5.40: Smith chart showing the measured input impedance of two bar-shaped boosters sized $3.2 \times 3.2 \times 10 \text{mm}^3$ each, with a 3.3mm gap, with a matching network only comprising $0 \Omega$ resistors

Instead, with two boosters, the matching circuits constructed in a ladder topology, may employ a series inductor of just $36 \mu \text{H}$. Two networks are built for the LTE700 band: the first one having the response from Figs. 5.41 and 5.42, and the second one with the response as in Figs. 5.43 and 5.44.

Based on the Smith chart diagrams, testing the radiating systems with other series inductor values for canceling the reactive part of the impedance is deemed unnecessary because, contrarily to what happened in the low frequency region, the upper and lower frequencies of the LTE700 band are quite even in the impedance loop. Furthermore, this balance in the position is also seen in the rectangular graphs with the $S_{11}$ curves having similar but opposed trends at the edges of the band. This would not be so if the loop were not centered even if, as mentioned, the two edge-most frequencies were in similar position with respect to the loop as happens herein.

The second matching network outperforms the first one: even though the peak antenna efficiency is five points in percentage less, the overall behavior of the radiating system within the whole band is much more uniform. On top of that, the average efficiency is slightly larger. Even in the hypothetical case of a lower antenna efficiency but a response as peaky as in Fig. 5.42, depending on the difference in efficiency, the more uniform one would still be the better radiating system.

With these results, the 10mm-long bar-shaped booster is considered to perform well
Figure 5.41: Smith chart showing the measured input impedance, in the LTE700 band, of the two bar-shaped boosters sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.10

Figure 5.42: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the LTE700 band, of the two bar-shaped boosters sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.10. Average antenna efficiency $\eta_{a,avg} = 56\%$
5.4. Measured Performance of Booster Candidate

Figure 5.43: Smith chart showing the measured input impedance, in the LTE700 band, of the two bar-shaped boosters sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.11

Figure 5.44: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the LTE700 band, of the two bar-shaped boosters sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.11. Average antenna efficiency $\eta_{a,avg} = 58\%$
in all the three frequency regions in single-port configurations. The following sub-section is devoted to measuring the performance when the candidate booster has to operate in a triple-port configuration, so having all four required boosters in place.

### 5.4.5. Triple-Port Configuration

The definitive test for the bar-shaped booster comes with the triple-port printed circuit board: it is mandatory that the radiating system performs well in all three frequency regions switching between the signals from the various ports. In a real scenario, a mobile phone should be able to use any of the three main cellular generation standards—2G, 3G or 4G—at any time.

A prototype allocating boosters in each of the three pads is built and the best-performing matching networks from the single-port setups are installed as seen in Figs. 5.45 and 5.46.

Figure 5.45: Photo of matching network close-ups from the final triple-port PCB; (left) high frequency region; (middle) low frequency region; (right) LTE700 band

As the low frequency region and LTE700 band are proximate in frequency, diametrically opposed boosters pads are chosen so as to increase the distance between them and, hopefully, the isolation between the two ports as well. The remaining booster pad, which shares a longer edge of the ground plane with the LTE booster—because even with two boosters together, they act as a single booster,—is reserved for the high frequency region.

Due to the necessity of having all the boosters in place and matched before taking any measurements, the matching networks from the single-port configurations are replicated; this ensures certain level of impedance matching that may be adjusted a posteriori based on the resulting $S_{11}$. Therefore, the chosen circuits are the ones whose schematics correspond to Figs. B.2 (low frequency region), B.9 (high frequency region), and B.11 (LTE700 band.)

Having the PCB completely prepared, the reflection coefficient for the low frequency region is analyzed for adjusting some components of the broadband matching circuit. After doing so, the performance of the booster in the 824-960MHz frequency range while the remaining ones are loaded with 50Ω is as seen in Figs. 5.47 and 5.48.

On top of the unbalanced reflection coefficient resulting in a better match at the higher
5.4. Measured Performance of Booster Candidate

Figure 5.46: Photo of the final triple-port PCB including four boosters sized 3.2x3.2x10mm³ each; the PCB is made of 1mm-thick FR4 with \( \varepsilon_r = 4.15 \) and loss tangent \( \tan \delta = 0.013 \); the ground plane has 120x60mm² dimensions.

Figure 5.47: Smith chart showing the measured input impedance, in the low frequency region, of the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, using the matching network from Fig. B.12 in a triple-port configuration; the high frequency region and LTE700 band boosters are loaded with 50Ω.
frequencies, there are certain issues in the lower frequencies. The reduction in antenna efficiency at the 824MHz part is believed to be also caused by mutual couplings with the LTE700 port, which is yet to be seen when said port is measured.

Anyway, the overall performance is good with an average efficiency of 55%, a match of at least -8dB within the whole band, just with the inconvenience of a slightly less than 40% efficiency in a small part of the low frequency region. For the most part, the $\eta_a$ is at least 50%.

The L-topology circuit is tested next with the focus of interest on the high frequency region. After minor adjustments, the measurements as taken are represented in Figs. 5.49 and 5.50.

Surprisingly, the input impedance does not behave as in the single-port scenario; even though the other ports are loaded with 50Ω, isolation issues still are not expected due to the huge differences in frequency with the low-frequency ports. Although not represented here, the $S_{12}$ measurements taken between the high frequency region and the other two boosters—one at a time—have shown isolations better than 30dB. Therefore, a possible explanation for this issue is that the components in the other matching networks being connected to the ground plane have an impact on the impedance of the booster for the high frequency region.
5.4. Measured Performance of Booster Candidate

Figure 5.49: Smith chart showing the measured input impedance, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.9 in a triple-port configuration; the low frequency region and LTE700 band boosters are loaded with 50Ω

Figure 5.50: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.9 in a triple-port configuration; the low frequency region and LTE700 band boosters are loaded with 50Ω. Average antenna efficiency $\eta_{a,avg} = 74\%$
Chapter 5. Ground Plane Boosters with a Bar Form Factor

Albeit the response is not bad, it is convenient to make the response more uniform within the whole band, even if the bandwidth is about 45%. With the L-topology, the 1.71GHz part suffered an increase in reflection coefficient that translated into a worse antenna efficiency around that frequency. So, going back to a T-topology, aided by Microwave Office®, the performance of the booster is improved as shown in Figs. 5.51 and 5.52.

Figure 5.51: Smith chart showing the measured input impedance, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, using the matching network from Fig. B.13 in a triple-port configuration; the low frequency region and LTE700 band boosters are loaded with 50Ω.

Not only has the antenna efficiency increased in average, but it has become much more constant within the band, being at least 70% and even 75% in a rather wide part. Moreover, the reflection coefficient is better than -10dB too. In conclusion, the addition of the third component has improved the performance significantly. It is interesting to point out that the series inductor is 3.7nH (Fig. B.13), whereas in the single-port case it was just 1.5nH (Fig. B.6), which was an appropriate argument for considering its removal from the circuit in the first place.

Turning now to the LTE700 band, the two bar-shaped boosters have been put together and matched with a network in a ladder topology. After finely adjusting the components being used, the LTE700 booster behaves as in Figs. 5.53 and 5.54.

The fact that the upper frequencies experience a situation similar to the booster for the low frequency region is very apparent. It is convenient to measure the $S_{12}$ parameter (Fig. 5.55) in order to find out whether the problem is precisely the isolation between the two ports.
5.4. Measured Performance of Booster Candidate

Figure 5.52: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the high frequency region, of the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, using the matching network from Fig. B.13 in a triple-port configuration; the low frequency region and LTE700 band boosters are loaded with 50$\Omega$. Average antenna efficiency $\eta_{a,avg} = 77\%$

Figure 5.53: Smith chart showing the measured input impedance, in the LTE700 band, of the two bar-shaped boosters sized 3.2x3.2x10mm$^3$ each, with a 3.3mm gap, using the matching network from Fig. B.14 in a triple-port configuration; the low and high frequency region boosters are loaded with 50$\Omega$. 
Figure 5.54: Rectangular chart showing the measured $S_{11}$ parameter and antenna efficiency $\eta_a$, in the LTE700 band, of the two bar-shaped boosters sized $3.2 \times 3.2 \times 10\text{mm}^3$ each, with a $3.3\text{mm}$ gap, using the matching network from Fig. B.14 in a triple-port configuration; the low and high frequency region boosters are loaded with $50\Omega$. Average antenna efficiency $\eta_{a,\text{avg}} = 54\%$

Figure 5.55: Rectangular chart showing the measured $S_{12}$ parameter for the LTE700 and low frequency region bar-shaped boosters sized $3.2 \times 3.2 \times 10\text{mm}^3$ each, with a $3.3\text{mm}$ gap, using the matching networks from Figs. B.14 and B.12, respectively, in a triple-port configuration; the high frequency region booster is loaded with $50\Omega$. 


The previous diagram makes it clear that the isolation has an effect on the performance in the two bands. The performance at both ports is considered to be quite satisfactory though; both have a good impedance matching and the worst antenna efficiency is around 40%, quite good values that for cellular services operating with such large wavelengths. In average, the efficiency is almost 55% and most of the frequencies in-band are above the 50% mark.

Besides, one last measurement is presented. It remained unknown whether a radiating system comprising bar-shaped boosters would feature an omnidirectional radiation pattern. Initially, without any measurements in this regard, the thought was that the radiation diagrams would indeed be quite omnidirectional as, once again, the main radiator of the whole structure is the ground plane. In the past, it has been seen that such structures radiate in all directions in a relatively similar way, and this has proven right for the current triple-port prototype as well as seen in Fig. 5.56 where the main 2D cuts have been plotted at several frequencies.

Figure 5.56: Main measured 2D radiation pattern cuts for the three ports of the final prototype including bar-shaped boosters sized 3.2x3.2x10mm³ each and respectively matched in impedance

Therefore, with this last measurement, it can be concluded that a bar-shaped booster with a 3.2x3.2x10mm³ size provides operation with overall good performance in all the three frequency regions in a platform like a smartphone.
5.5. Conclusions

The complete design and development processes of a demonstrator for a ground plane booster featuring a bar shape have been carried out in this chapter.

In the initial stages of the study, two booster candidates were presented, one of which was finally prototyped for performance testing purposes. In simulations and on prototypes, several matching networks were tested in pursuit of maximizing the transducer power gain and, ultimately, the antenna efficiency of a radiating system comprising said booster.

In the end, the measurements taken on the final prototype of this demonstrator have been positive, concluding that the RF performance is good overall. Additionally, even though the prototype used micro coaxial cables, its results have been compared to the cube-shaped ground plane booster product developed by Fractus® [1], whose evaluation boards use U.FL cables that are characterized by larger losses than micro coaxial cables. The comparison is shown in Table 5.1.

<table>
<thead>
<tr>
<th>Ground Plane Booster</th>
<th>5x5x5mm³ (U.FL cable)</th>
<th>3.2x3.2x10mm³ (micro coaxial cable)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Volume</td>
<td>125mm³</td>
<td>102mm³</td>
</tr>
<tr>
<td>Footprint (without gap)</td>
<td>25mm²</td>
<td>32mm²</td>
</tr>
<tr>
<td>Footprint (with gap)</td>
<td>32.5mm²</td>
<td>65mm²</td>
</tr>
<tr>
<td>LTE700 (min-max $\eta_a$)</td>
<td>40% - 59%</td>
<td>40% - 64%</td>
</tr>
<tr>
<td>LTE700 (average $\eta_a$)</td>
<td>50%</td>
<td>54%</td>
</tr>
<tr>
<td>LTE700 (min-max gain)</td>
<td>-1.5dBi - 0.6dBi</td>
<td>-1.6dBi - 0dBi</td>
</tr>
<tr>
<td>LFR (min-max $\eta_a$)</td>
<td>31% - 59%</td>
<td>38% - 63%</td>
</tr>
<tr>
<td>LFR (average $\eta_a$)</td>
<td>53%</td>
<td>55%</td>
</tr>
<tr>
<td>LFR (min-max gain)</td>
<td>-2.1dBi - 0.6dBi</td>
<td>-2.4dBi - 0.5dBi</td>
</tr>
<tr>
<td>HFR (min-max $\eta_a$)</td>
<td>59% - 81%</td>
<td>69% - 83%</td>
</tr>
<tr>
<td>HFR (average $\eta_a$)</td>
<td>72%</td>
<td>77%</td>
</tr>
<tr>
<td>HFR (min-max gain)</td>
<td>1.7dBi - 3.2dBi</td>
<td>2.6dBi - 4.2dBi</td>
</tr>
</tbody>
</table>

Table 5.1: Comparison of the product cube-shaped ground plane booster [1] in a 3-port evaluation board using U.FL cables and the bar-shaped ground plane booster in a 3-port prototype using micro coaxial cables

A rule of thumb regarding the cable differentiation is that antenna efficiencies with micro coaxial cables are, on average, on the order of 5% better than with U.FL cables. So, even if the table shows better antenna efficiencies for the bar-shaped booster, the results would probably be very similar if both boosters were to be tested in the same conditions.

With this comparison this chapter concludes having defined a booster with a bar form factor characterized by a height of 3.2mm. It has been demonstrated that boosters with the aforementioned bar shape feature a good RF performance in the three frequency regions of interest, namely the LTE700 band, the low frequency region and the high frequency region.
Chapter 6

Conclusions

A ground plane booster with a form factor fulfilling the 3.2mm-height requirement has been satisfactorily designed and prototyped, which is suitable for the new emergent smartphone devices with slim form factors. This booster has been made as a demonstrator for a possible future booster product together with the already-existing cube-shaped ground plane booster [1].

Altogether, it has been seen that a bar-shaped booster is feasible and performs well in radiating structures comprising a ground plane with the typical size of a smartphone ground plane. The final comparison has shown that the RF performance of the designed booster is very similar to that of the cubic booster, which in fact has been the target to achieve in this regard despite having a minor volume of material but a greater footprint.

During the design of a booster with a bar form factor, the implications of each of its dimensions have been explored. The knowledge obtained from the simulations focused on sweeping each dimension independently has aided in finding the size and shape of the ultimate booster candidate thus avoiding making endless trial-and-error experiments that would have taken much more time. In addition, this knowledge may be used again in the future if other form factors are to be designed. It is also noteworthy that, in this particular case, the existence of a ground plane booster that proved to feature a high RF performance has been a convenient guidance for setting the dimensions of a new booster so as to reach inherent bandwidth values that would be suitable.

A proposed future line of work is the combination of various boosters with a bar form factor in a single port, for instance using transmission lines and junctions, and provide impedance matching in different frequency regions (at least two, like the low and high frequency regions) owing to a single matching network specially tailored for this purpose. Such 1-port solution has been done for cubic boosters as related in [34]. This kind of solution may be convenient for circuitry and chips including one input port due to already-installed diplexers that separate the frequency bands within the chips. It may be advantageous as
well for reducing the number of lumped components in the network.

With respect to the proposed solution, it may be convenient to further study the effect that the booster position has on the ground plane mode excitation. Moreover, there are many matching network topologies which could possibly exploit the potential of the booster technology even more. For such improvement, it is advisable to attempt simulations with the most accurate possible behavior of the lumped components and other phenomena such as the effect of the pads or the presence of the other boosters. In this regard, alterations in the reflection coefficient due to possible mutual couplings could be studied as well.

A last possible line of work is developing additional form factors based on different footprints or volumes available for installing boosters in the device that, otherwise, would deny using a bar-shaped booster with the dimensions defined in this thesis.

Overall, making this thesis has produced three major results: the assessment of the feasibility of a new product, further knowledge in the field of ground plane boosters, wherein the most relevant aspect is the existence of optimum width and gap pairs of values for each possible depth dimension, and a new patent application.

Said knowledge consists on the maximization of the inherent bandwidth without maximizing the volume of material, which is not apparent.

And finally, regarding the feasibility of a new product, as a result of the work done in this master thesis, a new ground plane booster has been developed at *Fractus®* with the proposed dimensions of 3.2x3.2x10mm$^3$ [35] as seen in Fig. 6.1.

![Figure 6.1: (left) Photo of booster products with cube form factor and bar form factor, both designed by the R&D Fractus® Department [1][35]; (right) Photo of booster products with bar form factor designed by the R&D Fractus® Department [35]](image-url)
Bibliography

[1] “Fractus mXTEND” Antenna Booster,” *Fractus*®, FR01-S4-250


[35] “Fractus BAR mXTEND” Antenna Booster,” *Fractus®, FR01-S4-232*
Glossary

**BB-MN**  Broadband Matching Network

**BW**  Bandwidth

**CDMA**  Code Division Multiple Access

**ESR**  Equivalent Series Resistance

**GSM**  Global System for Mobile

**HFR**  High Frequency Region (1710MHz–2690MHz)

**LFR**  Low Frequency Region (824MHz–960MHz)

**L-MN**  L-Topology Matching Network

**LTE**  Long Term Evolution

**LTE700**  Long Term Evolution 700 Band (698MHz–798MHz)

**MN**  Matching Network

**PCB**  Printed Circuit Board

**RF**  Radiofrequency

**SMD**  Surface Mount Device

**SWR**  Standing Wave Ratio

**T-MN**  T-Topology Matching Network

**TPG**  Transducer Power Gain

**VSWR**  Voltage Standing Wave Ratio

**WCDMA**  Wideband Code Division Multiple Access
Appendix A

Matching Networks for Simulated Booster Candidates

A.1. Low Frequency Region

Figure A.1: Matching network for the simulated reference cube-shaped booster sized 5x5x5mm$^3$ and operating the low frequency region.

Figure A.2: Matching network for the simulated bar-shaped booster sized 3.2x3.2x7mm$^3$, with a 3.3mm gap, and operating the low frequency region.
A.2. High Frequency Region

Figure A.4: Matching network for the simulated reference cube-shaped booster sized 5x5x5mm$^3$ and operating the high frequency region

Figure A.5: Matching network for the simulated bar-shaped booster sized 3.2x3.2x7mm$^3$, with a 3.3mm gap, and operating the high frequency region
A.3. LTE700 Band

Figure A.6: Matching network for the simulated bar-shaped booster sized 3.2x3.2x10\,mm$^3$, with a 3.3mm gap, and operating the high frequency region

A.3. LTE700 Band

Figure A.7: Matching network for the two simulated reference cube-shaped boosters sized 5x5x5\,mm$^3$ each, and operating the LTE700 band

Figure A.8: Matching network for the two simulated bar-shaped boosters sized 3.2x3.2x7\,mm$^3$ each, with a 3.3mm gap, and operating the LTE700 band
Figure A.9: Matching network for the two simulated bar-shaped boosters sized 3.2x3.2x10mm$^3$ each, with a 3.3mm gap, and operating the LTE700 band.
Appendix B

Matching Networks for Measured Prototypes

B.1. Single-Port Configuration: Low Frequency Region

Figure B.1: First broadband matching network for a single-port prototype comprising the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, and operating the low frequency region. Results: Figs. 5.22 and 5.23
Figure B.2: Second broadband matching network for a single-port prototype comprising the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, and operating the low frequency region. Results: Figs. 5.24 and 5.25

Figure B.3: Third broadband matching network for a single-port prototype comprising the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, and operating the low frequency region. Results: Figs. 5.26 and 5.27

Figure B.4: Fourth broadband matching network for a single-port prototype comprising the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, and operating the low frequency region. Results: Figs. 5.28 and 5.29
B.2. Single-Port Configuration: High Frequency Region

Figure B.5: Fifth broadband matching network for a single-port prototype comprising the bar-shaped booster sized $3.2 \times 3.2 \times 10 \text{mm}^3$, with a 3.3mm gap, and operating the low frequency region. Results: Figs. 5.30 and 5.31

Figure B.6: T-topology matching network for a single-port prototype comprising the bar-shaped booster sized $3.2 \times 3.2 \times 10 \text{mm}^3$, with a 3.3mm gap, and operating the high frequency region. Results: Figs. 5.32 and 5.33

Figure B.7: First L-topology matching network for a single-port prototype comprising the bar-shaped booster sized $3.2 \times 3.2 \times 10 \text{mm}^3$, with a 3.3mm gap, and operating the high frequency region. Results: Figs. 5.34 and 5.35
Appendix B. Matching Networks for Measured Prototypes

B.3. Single-Port Configuration: LTE700 Band

Figure B.8: Second L-topology matching network for a single-port prototype comprising the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, and operating the high frequency region. Results: Figs. 5.36 and 5.37

Figure B.9: Third L-topology matching network for a single-port prototype comprising the bar-shaped booster sized 3.2x3.2x10mm$^3$, with a 3.3mm gap, and operating the high frequency region. Results: Figs. 5.38 and 5.39

Figure B.10: First ladder matching network for a single-port prototype comprising the two bar-shaped boosters sized 3.2x3.2x10mm$^3$ each, with a 3.3mm gap, and operating the LTE700 band. Results: Figs. 5.41 and 5.42
B.4. Triple-Port Configuration

Figure B.11: Second ladder matching network for a single-port prototype comprising the two bar-shaped boosters sized 3.2x3.2x10mm³ each, with a 3.3mm gap, and operating the LTE700 band. Results: Figs. 5.43 and 5.44

Figure B.12: Broadband matching network for the triple-port prototype comprising the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, and operating the low frequency region; $Q$ and $ESR$ (for capacitors only) parameters for central frequency indicated. Results: Figs. 5.47 and 5.48
Figure B.13: T-topology matching network for the triple-port prototype comprising the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, and operating the high frequency region; Q and ESR (for capacitors only) parameters for central frequency indicated. Results: Figs. 5.51 and 5.52

Figure B.14: Ladder matching network for the triple-port prototype comprising the bar-shaped booster sized 3.2x3.2x10mm³, with a 3.3mm gap, and operating the LTE700 band; Q and ESR (for capacitors only) parameters for central frequency indicated. Results: Figs. 5.53 and 5.54