Electromagnetic Susceptibility of Electro-Explosive Devices

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To my brother Jorge and to all the rest of heroes and victims of violence in Colombia.
# Table of Contents

Jury: ................................................................................................................................. ii

Financial and Technical Resources .................................................................................. iii

Chapter 1 Introduction ....................................................................................................... 2
  1.1 Research Topic .............................................................................................................. 2
  1.2 Project Objectives and Thesis Outline ......................................................................... 3

Chapter 2 Electromagnetic Susceptibility of Electro-Explosive Devices ......................... 5
  2.1 Modeling the Electromagnetic Susceptibility of EEDs ................................................ 5
    2.1.1 Initiation Mechanism ............................................................................................ 5
    2.1.2 EED with connection wires .................................................................................. 6
    2.1.3 Norton and Thévenin Equivalents ........................................................................ 7
    2.1.4 Antenna Theory .................................................................................................... 8
    2.1.5 Electric and Inductive Coupling ........................................................................... 9
    2.1.6 High Frequency Models ....................................................................................... 10
  2.2 High Frequency Electromagnetic Coupling Model .................................................... 11
  2.3 Testing the EM Response of Electro Explosive Devices ............................................. 14
    2.3.1 Destructive Tests .................................................................................................. 14
    2.3.2 Non-destructive Tests ......................................................................................... 15
  2.4 Conclusions ................................................................................................................ 16

Chapter 3 EM Susceptibility of EEDs to High Power Electromagnetic Excitations ............ 18
  3.1 Introduction .................................................................................................................. 18
  3.2 EED’s Transient Thermal Models ............................................................................... 19
    3.2.1 First-order Model ............................................................................................... 19
    3.2.2 Second-order Model ......................................................................................... 21
  3.3 Short Excitation .......................................................................................................... 22
  3.4 Repetition Gain .......................................................................................................... 22
  3.5 ANSYS Thermal Simulations ..................................................................................... 24
  3.6 Analytical Model Validation ...................................................................................... 25
  3.7 Electromagnetic Coupling Model .............................................................................. 27
    3.7.1 EED Ignition ...................................................................................................... 27
    3.7.2 Electromagnetic Model ..................................................................................... 27
  3.8 Assessment Against IEMI ......................................................................................... 28
<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>Acknowledgements</td>
<td>104</td>
</tr>
<tr>
<td>List of Publications</td>
<td>106</td>
</tr>
<tr>
<td>Conference Papers</td>
<td>106</td>
</tr>
<tr>
<td>Journal Papers</td>
<td>106</td>
</tr>
<tr>
<td>About the Author</td>
<td>107</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
</tr>
<tr>
<td>--------------</td>
<td>----------------------------------</td>
</tr>
<tr>
<td>AUT</td>
<td>Antenna Under Test</td>
</tr>
<tr>
<td>CDF</td>
<td>Cumulative Distribution Function</td>
</tr>
<tr>
<td>CW</td>
<td>Continuous Wave</td>
</tr>
<tr>
<td>DS</td>
<td>Damped Sinusoidal</td>
</tr>
<tr>
<td>DUT</td>
<td>Device Under Test</td>
</tr>
<tr>
<td>EED</td>
<td>Electro-Explosive Device</td>
</tr>
<tr>
<td>EM</td>
<td>Electromagnetic</td>
</tr>
<tr>
<td>EMC</td>
<td>Electromagnetic Compatibility</td>
</tr>
<tr>
<td>EMP</td>
<td>Electromagnetic Pulses</td>
</tr>
<tr>
<td>FP</td>
<td>Firing Probability</td>
</tr>
<tr>
<td>HPEM</td>
<td>High Power Electromagnetics</td>
</tr>
<tr>
<td>IED</td>
<td>Improvised Explosive Device</td>
</tr>
<tr>
<td>IEMI</td>
<td>Intentional Electromagnetic Interferences</td>
</tr>
<tr>
<td>LPM</td>
<td>Low Power Microwave</td>
</tr>
<tr>
<td>MCS</td>
<td>Monte Carlo Simulations</td>
</tr>
<tr>
<td>MEG</td>
<td>Mean Effective Gain</td>
</tr>
<tr>
<td>PDF</td>
<td>Probability Density Function</td>
</tr>
<tr>
<td>PRF</td>
<td>Pulse Repetition Frequency</td>
</tr>
<tr>
<td>RFI</td>
<td>Radio Frequency Interference</td>
</tr>
<tr>
<td>SABC</td>
<td>State Of The Art Blasting Cap</td>
</tr>
<tr>
<td>SF</td>
<td>Survivor Function</td>
</tr>
<tr>
<td>TDIE</td>
<td>Time Domain Integral Equation</td>
</tr>
<tr>
<td>VNA</td>
<td>Vector Network Analyzer</td>
</tr>
</tbody>
</table>
Chapter 1
Introduction

1.1 Research Topic

Electro-explosive Devices (EEDs) are electrical components commonly used in different applications nowadays. Their use extends from automotive industries, where they are used to activate the airbag system, to commercial mining and demolition applications [1]. Recently, these devices have been incorporated in ordnance, military equipment, and improvised explosive devices [2-4]. Since EEDs are electrically activated, they are susceptible to electromagnetic radiation. An unintended activation of this component could produce catastrophic consequences. For this reason, a study on the characteristics of the electromagnetic (EM) response of the EED immersed in each system is necessary to guarantee a safe operation.

The electromagnetic compatibility (EMC) determines the ability of an equipment or system to operate without generating intolerable electromagnetic disturbances and tolerating the disturbance of its electromagnetic environment. Electromagnetic susceptibility relates to the second part of this definition; it determines the degradation level produced in the equipment or system due to a determined EM disturbance. In the case of systems with EEDs, the electromagnetic susceptibility is related to an unintended ignition due to external EM fields and determines the disturbance characteristics, such as the waveform and the amplitude, that cause a failure. A study on electromagnetic susceptibility involves the characterization of the device or system EM response; that is, the capacity to receive EM fields and transmit the induced signals to critical components.

Waveforms of disturbing EM fields are diverse. EMC tests commonly assess the response to typical industrial or domestic electromagnetic environments according to the device’s use. In the last decades, High Power Electromagnetics (HPEM), which are signals with very intense EM fields, are being considered. HPEM can be generated by natural sources, such as lightning strikes, but also by man-made sources, such as nuclear bursts or radar systems [5]. In this thesis, the electromagnetic susceptibility of systems with EEDs in HPEM environments is studied.

A characteristic of systems with EEDs is the variability in the configuration of the connection wires. Each application that uses EEDs presents different characteristics. In mining, for example, long unshielded twisted pair wires are used to connect the firing circuit and the EED; while in ordnance, the firing circuit is usually close to the EED and both are inside metallic structures. In
addition, even in a specific application, the pattern in their geometry and length can change depending on the operation phase (i.e. storage, assembling, or service) and on the user requirements. The connection wires characteristics are determinant in the calculation of the electromagnetic coupling since they are the main input port of disturbances. Random characteristic in the system under study produce uncertainty in the EM susceptibility. This characteristic is specially addressed in this thesis.

1.2 Project Objectives and Thesis Outline

The main objective of this thesis is to study the radiated electromagnetic susceptibility of systems with EEDs. As a result, a model to determine the frequency response of the EED and the characteristics of the incident field able to more probably disturb the system is developed. To accomplish this task, the influence of the randomness in the wiring and the influence of different frequency dependent loads are studied. In addition, it is proposed the characterization of typical elements used in EED firing circuits and the study of the stochastic interaction between an incident field and wire configurations with random parameters. The influence on the susceptibility of the firing circuit variability is analyzed statistically, while the influence of the EED behavior in the coupling is analyzed deterministically.

Special interest is focused on the EM and thermal response of EEDs and the techniques of experimentally characterizing the high frequency response of the system’s elements. The main original findings obtained in the development of this work are listed below:

- Previous studies on the EM response of wired EEDs neglect the EED’s resonant behavior. In this thesis, an innovative high frequency model that includes the frequency dependence of the wires’ impedance and also the frequency behavior of the EED’s impedance is developed.

- This thesis is one of the first published studies that assess the EM susceptibility of EEDs against HPEM. Practical simplifications of the thermal response for short transient and CW excitations are formulated. Additionally, the characteristic of stacking energy in an EED is mathematically described by the Repetition Gain. Using numerical models, it is demonstrated that CW sources are the most disturbing signals in systems with EEDs.

- It is known that systems present a resonance region in which the electromagnetic response is maximized [6]. In this thesis, this concept is applied to a system with random properties. The resonance frequency, calculated as the frequency of optimal average coupling, is calculated for the case of an EED with a firing circuit of random
characteristics. This original contribution allows determining the frequency in which the system is more susceptible to incident EM fields.

- A system with random properties presents a random electromagnetic response. In this work, it is proposed to model the average behavior of a random system using an equivalent structure. It is shown that the average response of an EED with a random firing circuit can be replicated using an equivalent structure that has the system’s average effective volume.

- High-frequency experimental characterization of differential devices, such as an EED or its firing circuit, presents some challenges since instruments are designed to measure single-ended devices. In this work, a characterization technique to measure gain and impedance of differential devices in a wide frequency range is developed. For this, modern vector network analyzer’s capabilities of measuring mixed mode S-parameters are used.

This dissertation is organized as follows. A discussion of EEDs and methods to assess its EM susceptibility is presented in Chapter 2. Then, a study of the thermal effects of HPEM disturbances on a canonical EED is presented in Chapter 3. In Chapter 4, the susceptibility of EEDs against CW excitations is analyzed. Particularly, this chapter presents the frequency response of EEDs connected to wires with random geometries. Chapter 5 shows a technique for characterizing the EM response of differential devices, such as EEDs. The application of this technique for characterizing the firing circuit of Improvised Explosive Devices is shown in Chapter 6. Chapter 7 presents the model’s experimental validation using fiber-optic thermometry. Finally, general conclusions and remarks are presented in Chapter 7.
Chapter 2

Electromagnetic Susceptibility of Electro-Explosive Devices

The response to electromagnetic interference of systems with electro-explosive devices (EEDs) in their circuits has been a subject of interest in the last decades. Due to their initiation mechanism, EEDs can be fired by electromagnetic fields coupled in their connection wires. Therefore, different models to assess the radiated susceptibility of EEDs have been proposed. In this section, analytical and numerical models used to represent the coupling between electromagnetic fields and EED firing circuits are presented. In addition, current experimental techniques for assessing the susceptibility of EEDs are summarized.

2.1 Modeling the Electromagnetic Susceptibility of EEDs

2.1.1 Initiation Mechanism

Electro-explosive devices (EEDs) are commonly used in many industries (e.g. mining, automotive, and military) because of their versatility and simple operation mode [1]. According the initiation mechanism, there are three main types of EEDs: i) hot bridge-wire detonators, ii) exploding bridge-wire detonators, and iii) exploding foil initiators. The first two kinds are particularly susceptible to EM fields since they are low voltage devices [7]. One of the most common types of EEDs is the hot bridge-wire detonator, whose diagram is shown in Fig. 2.1. In this dissertation, we will only consider this type of detonator. Hot bridge-wire detonators are composed by a pair of feeding wires that connect a bridge-wire located inside a cylindrical metallic case. The bridge-wire is surrounded by a primary explosive that is activated when its temperature achieves a critical value. Normally, these devices are activated with DC current pulses or a capacitor discharge.

Hot wire-based EEDs have an all-fire current, which is the current that guarantees an activation with a 99% probability, typically in the order of 800 mA for low energy devices [8]. In addition, manufacturers also provide the no-fire current or the maximum no-fire threshold (MNFT), which is the current that guarantees a 0.1% firing probability with a 95% of confidence level [8]. The MNFT varies in each device, but it ranges from 100 mA for low energy EEDs to several Amperes for high energy devices [2]. One of the drawbacks of hot wire EEDs is that electromagnetic (EM) fields can
induce currents higher than the firing threshold in their connection wires; for this reason, these devices are susceptible to radio frequency interference (RFI).

An unintended activation can be achieved with different kinds of radiofrequency (RF) signals [2, 8]. In [9], the sensitivity of six types of EEDs against continuous wave (CW) and pulse modulated excitations is assessed and safe distance curves from the interference source are determined. Single and repetitive radiated pulses have been also considered. Although one short pulse could not induce enough current, the contribution of heating due to repetitive pulses could initiate the EED [2]. In that case, successive heating and cooling periods are presented in the EED’s bridge-wire [10]. On the other hand, the initiation due to a radiated single pulse is difficult to achieve because the bridge-wire heating presents a derivative behavior and the temperature shows an exponential response as a function of the time [11]. However, some wideband short pulses, such as the electromagnetic pulses (EMP) produced by nuclear detonations, can have enough energy to initiate EEDs. In [12], the effects high altitude electromagnetic pulses (HEMP) on EEDs are experimentally assessed using standard waveforms with rise time higher than 10 ns. It is concluded that EEDs can be initiated with pulses of energy higher than 0.1 mJ. In any case, the induced current mainly depends on the signal intensity, spectral content, and the characteristics of the firing circuit.

![Fig. 2.1. Schematic of a hot-wire detonator with typical dimensions and materials based on [13]](image)

### 2.1.2 EED with connection wires

A hot-wire based EED is composed by four parts: a pair of feeding wires, a bridge-wire, explosive charges, and a metallic case. Depending on the application and the manufacturer, dimensions and materials can vary. Fig. 2.1 shows the dimensions and materials of a typical EED based on the state-of-the-art (SABC) model proposed in [13]. From the EM standpoint, the EED and its firing circuit can be modelled by an antenna that receives the EM radiation and delivers the energy to the EED’s bridge-wire. Thus, the geometry and length of the connection wires mainly determine the capacity of the circuit to absorb incident radiated power.

Models used to study the EM susceptibility of EEDs usually consider the feed wires as the elements to which the EM field couples, and the EED as its load [8, 10, 14-18]. To determine the
characteristics of the electromagnetic field that causes the EED’s activation, this problem has been simplified representing the wires as standard antennas with known characteristics (e.g. dipoles and loops) [8, 10, 14, 15], applying analyses of worst case based on these canonical structures [8, 15-18], and assuming that the system is electrically short [19].

2.1.3 Norton and Thévenin Equivalents

Norton and Thévenin equivalents have been commonly employed to study the EM susceptibility of electric systems [20] and, particularly, EEDs. In [16], a Thévenin equivalent is used to represent the ignition circuit and a load impedance is used to represent the EED, as shown in Fig. 2.2. This equivalent circuit allows considering the ignition circuit as a receiving antenna. Therefore, the elements of this circuit are the open circuit induced voltage $V_A$, the antenna’s impedance $Z_A$ and the EED’s impedance $Z_{EED}$. For an antenna with gain $G$ and impedance’s real part $R_a$, the Thévenin voltage due to an incident CW electric field $|E|$ is [21]

$$V_A = \frac{|E| \lambda (R_a G)^{1/2}}{120},$$

(2.1)

where $\lambda$ is the wavelength. As a consequence, the power delivered to the EED yields

$$P_{EED} = \left(\frac{|E| \lambda}{\pi}\right)^2 \frac{G}{120} \frac{R_a R_{EED}}{(Z_A + Z_{EED})^2},$$

(2.2)

where $R_{EED}$ is the real part of $Z_{EED}$.

![Fig. 2.2. Thévenin equivalent circuit of an EED and its firing circuit.](image)

In [16], Einstein and Warner considered some assumptions to reduce the complexity of the problem. They considered devices in free-space and assumed impedance matching between the EED and the ignition circuit. Different standard structures (i.e. two-wire parallel, single wire, and small loop), with closed form expressions for the gain and the impedance, were used to represent the ignition circuit.

It is important to note that, although the equivalent circuits are practical to calculate the induced power in the EED, they are not accurate to calculate the scattered power by the ignition circuit in
general. In [21], it is concluded that Thévenin and Norton equivalents cannot predict structural scattering and that they only estimate the re-radiated power in special frequency ranges, depending if the antenna behaves as a series or a parallel RLC circuit.

2.1.4 Antenna Theory

Some studies on EED’s susceptibility have been developed using worst case analyses. Therefore, the device characteristics, such as the wiring geometry, the orientation, and the mismatch between the ignition circuit and the EED, are assumed to produce the maximum induced power in the EED. Considering these assumptions in (2.2), the maximum possible power induced due to an incident electric field can be calculated as

$$P_{Max} = \left( \frac{|E| \lambda}{4\pi} \right)^2 \frac{G}{30}$$  \hspace{1cm} (2.3)

This expression has been used to calculate safe field strengths for EED environments. In [8], the ignition circuit is modeled using a half-wave resonant dipole; then, the gain in (2.3) is taken as $G = 1.64$ dB. Since these calculations were compared with measurements in offshore oil platforms, a 6 dB factor was incorporated to consider the field enhancement due to nearby metallic structures. Results showed that the model provides maximum limits in the induced current for VHF and UHF bands; however, it is not valid for MF and HF bands.

In [14], maximum current curves as a function of the frequency due to a 1 V/m incident field are calculated. The firing circuit is modeled using dipole and loop matched antennas and the EED is modeled as a 0.9 Ω resistive load. The matching between the antenna and the EED is obtained by letting to the firing circuit works partially as antenna and as transmission line. Matching of the reactive components can be achieved for wire lengths lower that $\lambda/2$, where the capacitive impedance of the antenna can be tuned by the inductive impedance of the parallel-wires transmission line. The calculated curves were experimentally verified in the typical operation environment of a surface mine, obtaining induced currents lower than 36 mA for a 1 V/m incident field. Dipole and loop configurations and wire lengths ranging from 3 m to 800 m were considered.

Similar worst-case approaches are currently used to calculate safe power densities and distances of transmitters in the vicinity of EEDs. In Fig. 2.3, for example, two curves of safe power density for munitions with EEDs in normal service configuration and damaged or disassembled are shown. These curves show the canonical “band-pass filter” response, which is a general behavior that typical systems present [6].
2.1.5 Electric and Inductive Coupling

Other method to study low frequency induced currents in EEDs is by means of electric coupling and magnetic induction models. In [17], an electric field coupling model and a magnetic induction model are presented to calculate the effect of EM fields produced by power transmission lines on EEDs connected to open ended wires and loops, respectively. The expressions used in that work to calculate the short-circuit induced current for the electric and magnetic couplings are

\[ |I_{sc}| = \omega \varepsilon_0 |E| S \quad (2.4) \]
\[ |I_{sc}| = \omega \mu_0 |H| S / R_{EED} \quad (2.5) \]

where \( S \) is the surface area of the wiring (e.g. loop area), \( \omega \) is the excitation frequency, \( R_{EED} \) is the EED’s resistance, and \( |E| \) and \( |H| \) are the electric and magnetic field strengths, respectively.

Similar analyses, but in time domain, have been developed to study the effect of pulsed excitations on EEDs [22, 23]. In [22], where an Electromagnetic Pulse (EMP) excitation is considered, the wiring is represented by a dipole and a closed loop oriented for maximum pickup. The expressions formulated to calculate the open circuit induced voltage for both cases are

\[ V_A(t) = \frac{L}{2} \left[ 1 + \left( \frac{L}{\lambda} \right)^2 \right] E(t) \quad (2.6) \]
\[ V_A(t) = \mu_0 A \partial H(t) / \partial t \quad (2.7) \]
where $\lambda$ is the wavelength, $L$ is the dipole’s length, and $A$ is the loop’s area.

From (2.7), the induced voltage for a continuous wave (CW) excitation on a short loop can be calculated as

$$|V_A| = \mu_0 A \omega |H| \cos \gamma,$$

(2.8)

where $\gamma$ is the angle formed by the magnetic field vector and the normal vector of the loop’s plane. Similarly, the open–circuit voltage for a short dipole is given by [19]

$$|V_A| = l |E| \cos \gamma,$$

(2.9)

where $l$ is the dipole’s length and $\gamma$ is the angle between the electric field vector and the dipole axis. This approach, together with a Thévenin equivalent model, is used in [19] to study the influence of the frequency and the dimensions on the coupling. In addition, calculations are compared with results of destructive tests. Particularly, a system’s resonant frequency of 400 MHz is verified for an EED with 5 cm wires.

**2.1.6 High Frequency Models**

Most of the models shown above represent the EED as a pure resistive element with a constant, low resistance. However, EEDs present a resonant behavior as the frequency increases [13]. Thus, this frequency dependence and, as a consequence, the variable impedance matching between the connection wires and the EED becomes an important factor to be taken into account in the susceptibility analysis.

An analytical model of the EED’s input impedance as a function of the frequency was presented by Lambrecht in [13]. This model represents the EED with three transmission lines in cascade loaded with the bridge wire. The type of each transmission line depends on the structure and materials in each transversal section, as shown in Fig. 2.1. Thus, the first section, at the EED input, can be represented by a two-wire transmission line in air, the second, which corresponds to the insulating header, by a twin-axial line in rubber, and the third, the primary explosive section, by another twin-axial line in lead-azide. Knowing the characteristic impedance of the transmission lines, the input impedance of each section can be calculated as [24]

$$Z_{in_k} = \frac{Z_{L_k} + j Z_{0_k} \tan \theta_k}{Z_{0_k} + j Z_{L_k} \tan \theta_k},$$

(2.10)

where $k=1,2,3$ is the section’s number, $Z_{0_k}$ is the section’s characteristic impedance, $Z_{L_k}$ is the load impedance, and $\theta_k = (2\pi f / v_{p_k}) \Delta l_k$ is the section’s electrical length. Thus, the load of the last section corresponds to the bridge wire, and the input impedance of the two-wire line section
corresponds to the EED’s input impedance. The characteristics of the transmission lines used to represent a typical EED are summarized in Table 2.1.

In [15], a full-wave electromagnetic simulation is used to calculate the input impedance of an EED at 900 MHz. This impedance is introduced in a Thévenin equivalent circuit to predict the coupling between a transmitter with a half-wave dipole and an EED with parallel wires in free space. The results are compared with measurements, obtaining a predicted power coupling of -69.2 dB at a distance of 10 cm.

Table 2.1. Transmission line parameters of the EED model. Adapted from [13].

<table>
<thead>
<tr>
<th>#</th>
<th>Section</th>
<th>Transmission Line</th>
<th>( \Delta l ) (mm)</th>
<th>( \varepsilon_r )</th>
<th>Per unit length parameters</th>
<th>( Z_0 ) (Ω)</th>
<th>( \nu_p/c )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Feed wires</td>
<td>Two-wire</td>
<td>7.9</td>
<td>1</td>
<td>12.13</td>
<td>0.918</td>
<td>275.10</td>
</tr>
<tr>
<td>2</td>
<td>Insulating Header</td>
<td>Twin-axial in rubber</td>
<td>37</td>
<td>3</td>
<td>49.90</td>
<td>0.892</td>
<td>133.70</td>
</tr>
<tr>
<td>3</td>
<td>Primary Charge</td>
<td>Twin-axial in lead-azide</td>
<td>3.3</td>
<td>17</td>
<td>212.07</td>
<td>0.892</td>
<td>64.85</td>
</tr>
</tbody>
</table>

Recently, Mora et al. proposed a methodology to calculate the transfer function between a short external excitation (i.e. impinging EM field) and the EED’s bridge-wire [25]. This study included the mathematical description of the electromagnetic coupling with the wiring, the power transmission inside EED (i.e. from the input to the bridge-wire), and the transient energy conversion into heat in the bridge-wire.

### 2.2 High Frequency Electromagnetic Coupling Model

In this section, a high frequency model to calculate the induced power in an EED is presented. In a differential mode coupling, the connection wires act as receiving antenna and the incident electric field induces currents that feed the EED. Thus, the system (i.e. EED and wires) can be represented by an equivalent circuit, as shown in Fig. 2.4, in which the EED is represented by a load impedance \( Z_{EED} \) and the wires are represented by a Norton equivalent current source \( I_N \) with an internal impedance \( Z_A \). Note that the dipole shown in Fig. 2.4a is only for illustrative purposes.

![Diagram](image1.png)

**Fig. 2.4 a)** Diagram and **b)** equivalent circuit of an incident plane wave on an EED with connection wires. Typical sections of a hot-wire detonator are depicted.
The wires’ parameters (i.e. Norton source and the source equivalent impedance) can be calculated using analytical expressions or numerical methods. The input impedance of the EED can be calculated using (2.10). Fig. 2.5 shows, as an example, the EED’s input impedance as a function of the frequency compared with the simulated input impedance of a 21.3-cm long dipole with a wire diameter of 0.7 mm. The EED presents low impedance at low frequency because, in this range, the impedance value is close to the bridge-wire resistance at DC, which is about 1.3 Ω. In contrast, the wires, with impedance \( Z_A \), show high impedance at low frequency due to its open circuit termination. Most connection circuits of EEDs before its activation are open-ended; therefore, the response presented in Fig. 2.5, can be considered as typical in most practical cases. The figure also shows the resonant behavior of both structures, with alternating phase and magnitude as the frequency increases. As a result, there are multiple frequencies in which the EED’s impedance is close to the conjugate complex of the wires’ impedance. When this condition is satisfied, the maximum power transfer is obtained in the circuit of Fig. 2.4b. The first two frequencies that comply with this condition are 550MHz and 900MHz, as depicted in Fig. 2.5.

![Fig. 2.5 Frequency response of the a) magnitude and b) phase of the input impedances of the EED and of the 21.3-cm long wires in dipole configuration.](image)

Fig. 2.6 presents the induced current in the wires due to a 1 V/m parallel incident electric field. It shows two maximum values in the observed frequency range, which correspond to the dipole’s resonance frequencies. Using the Norton equivalent, the power delivered to the EED \( P_d \) can be obtained using the following expression [21]:

\[
P_d = \frac{1}{2} |I_N|^2 \left| \frac{Z_A}{Z_A + Z_{EED}} \right|^2 \text{Re}[Z_{EED}],
\]

where \( I_N \) is the Norton current, \( Z_A \) is the wires’ impedance, and \( Z_{EED} \) is the EED’s input impedance.
To verify experimentally the coupling model, the absorbed power in a 100 Ω load connected to the wires of the previous example was measured in the anechoic chamber of the Universidad de los Andes. The incident plane wave was generated connecting a reference antenna to a port of a vector network analyzer (VNA) with a wideband amplifier. Two reference antennas (Log Periodic Dipole Array LPDA and Double-Ridged Guide Antenna HORN) were used to cover the frequency range between 200MHz and 3GHz. The received power in a balanced port of the VNA was obtained following the procedure presented in [26] and detailed in Chapter 5. The calculated and measured powers are presented Fig. 2.7a, showing a good agreement.

When the same wires are loaded with the EED, its impedance dependence with the frequency changes the frequency response in the absorbed power. This was calculated using (2.11) and is presented in Fig. 2.7b. As expected, the power presents local maxima at the frequencies in which the EED and the wires impedances, presented in Fig. 2.5, are matched. In addition, it is possible to see that the global maximum value is obtained in the first matching frequency. That is the effect of
the short circuit current frequency response; the power follows the current general tendency of decreasing its magnitude with the frequency.

2.3 Testing the EM Response of Electro Explosive Devices

In addition to mathematical models, experimental tests have been used to determine the EM response of EEDs. In this section, some of these experimental techniques are described.

2.3.1 Destructive Tests

Due to the difficulty of measuring the induced power in EEDs without modifying their electromagnetic and thermal properties, destructive tests have been commonly used to study the response of EEDs to radiated EM fields.

A first experimental study on the response of EEDs to RF excitations was performed by Thompson [9]. In that work, the CW signals in the range of 1.5 MHz to 900 MHz were directly injected in different kinds of EEDs in differential (pin to pin) and common (pin to case) modes. In differential mode, the power required to fire the EEDs was similar to the required one using a DC signal. Ignitions with power levels as low as 77 mW were reported. In common mode, firing power levels resulted 10 times higher than for DC excitations for most of the cases. Similar tests for common [27] and differential mode [19] are reported. In [19], EEDs with 5 cm wires were excited with CW signals in the range of 154 MHz to 900 MHz and between 0.8 and 1.2 W. Ignitions at 400 and 600 MHz were achieved with power levels of 0.9 W.

Tests with pulsed signals have been also reported. In [23], single pulses and pulse trains were used. That study showed that the EED presents an adiabatic heating when it is fed with single short pulses (i.e. pulse width < 8 ms). In this case, most of the supplied energy is dissipated in the bridge-wire; then, the ignition is achieved when the energy exceeds a threshold. In [23], a threshold of 8 mJ was obtained. In addition, the study showed that in a pulse train excitation, EEDs present heating and cooling cycles; for this reason, the activation energy is higher than for single pulse excitation. In [28], the EED’s activation due to CW pulses was experimentally verified. With a 3.07 GHz signal, a 50% firing average power of 0.187 W was obtained. The effects of other kinds of pulsed signals, such as HEMP, have been also studied using destructive tests [12].

Remote activation tests using electromagnetic fields have been also performed. In [29], a radiation test was performed on an antipersonnel mine that included an EED as part of its ignition mechanism. The mine was illuminated with a 50 V/m CW electric field in the range of 300 MHz to 1 GHz. Different polarizations and incidence angles were used. The objective of this test was to
compare the measured mine’s resonance frequencies with full-wave simulations. The obtained resonance frequencies, determined as the frequencies in which a detonation were achieved, were close to 543, 692, and 893 MHz.

2.3.2 Non-destructive Tests

Different non-destructive tests to quantify the coupling between an incident EM field and the EED’s final load (i.e. the bridge-wire) have been proposed. This kind of measurement presents a challenge because, as mentioned earlier, the instrumentation should not affect neither the electrical nor the thermal behavior of the system. Initially, a thin film thermocouple in the vicinity of the bridge-wire was proposed to quantify the effects of RF pick up on EEDs [30]. Other indirect techniques consist of measuring the current in the wires with a RF current transformer or calculating the bridge-wire current by indirect measurements with a voltage divider [31]. However, outputs of these measurements are disturbed by EM fields and the conductive elements, required for these techniques, affect the EM near fields of the device under test.

One alternative to avoid this drawback is the use of fiber-optic based sensors to measure the bridge-wire dissipated power from its temperature increase [32, 33]. In [34], a method based on infrared radiation thermometer is proposed to determine safety distances in HPEM environments. Specifically, the EED response to a damped sinusoidal (DS) excitation is shown in that study. In [35], a fiber-optic sensor based on Fabry-Perot Interferometry is used to verify the extrapolation of the temperature measurement due to higher electromagnetic field levels at specified frequencies. The test was performed in a shielded enclosure in the range of 400 MHz to 8 GHz and with field strengths between 20 V/m and 160 V/m. A similar probe was used in [36], in which the susceptibility of EEDs connected to short wires (<25 cm) in dipole and parallel configuration was tested for CW in free space. Maximum bridge wire currents of 170 mA and 77 mA were reported for dipole and parallel configurations, respectively, with a 50 V/m incident field. Table 2.2 summarizes the characteristics of a list of tests that use radiated RF signals to excite EEDs in different setups.

Currently, fiber optic sensors, based on the temperature-dependent band-gap of a GaAs crystal [37] and fluoroptic thermometry [38], have been adapted for this application and are available commercially. The disadvantage of fiber-optic based measurements is that only modified EEDs can be measured since the explosive content of the EED should be extracted to insert the sensor. A solution to this issue is to packing the EED with an inert material with electromagnetic and thermal properties similar to those of original fillers [39].
Another approach to characterize the electromagnetic response of EEDs is by measuring their impedance as a function of the frequency. As mentioned earlier, modeling the EED impedance as a low resistive load (~1 Ω) is not valid for RF frequencies. The frequency dependence of EEDs was modeled with the SABC model, explained in Section 2.1.6 and proposed in [13]. To represent the typical electromagnetic behavior of an EED, this model was fitted from differential mode measurements of the input impedance of 108 EEDs in the range of 10 MHz to 6 GHz. The characterization was performed using a vector network analyzer (VNA) and a test set to measure S-parameters of two-port devices.

### 2.4 Conclusions

This chapter presented different techniques reported in the literature to study the electromagnetic susceptibility of EEDs. Initially, worst case analyses based on simplified circuit equivalent models were formulated. New models have been proposed to increase the accuracy and to predict the effects of different types of signals. Although CW excitations have been mainly considered, currently, high power pulsed signals, such as damped sinusoidal waveforms, are been studied. Recent studies also show the importance of the resonant behavior of these devices. They have
shown that excitations tuned at the system resonant frequencies induce higher currents and can cause remote detonations.
Chapter 3

EM Susceptibility of EEDs to High Power Electromagnetic Excitations*

This chapter presents an analysis of the thermal response of a hot wire electro-explosive device (EED) excited with different transient signals. First-order and second-order analytical models to calculate the thermal response of an EED are assessed taking as reference numerical simulations obtained using ANSYS®. Appropriated thermal models for early-time and late-time responses are deduced from this analysis. The models are used to assess the electromagnetic susceptibility of a wired EED for different electromagnetic pulsed environments. Radiated signals produced by a mesoband radiator, two types of radars, and a hyperband radiator are considered. The radar signal proved to be the most disturbing source because of its highest duty-cycle and its flat spectral response around a specific frequency. It is shown that even the temperature firing threshold can be exceeded with the radiated field produced by a radar of 200kW of output power located at a distance of 5 meters.

3.1 Introduction

One of the most common type of EED is the hot-wire detonator, which is described in Chapter 2. An ignition is achieved when the bridge-wire temperature reaches a critical value. Normally, these devices are activated with DC currents or a capacitor discharge. However, RF currents with different time-dependent waveforms can also produce enough dissipation power for activation. These signals can be induced when the EED is exposed to an incident electromagnetic field [10, 34]. Environments with pulsed and damped-sinusoid signals have been experimentally and theoretically considered in previous studies [10, 34, 40, 41].

Approximate thermal EED models are used in the literature. The Rosenthal first-order model [11], which is based in the resulting experimental temperature waveform, has been commonly used. A second order model, the fitted wire model, was also proposed in [42]. These models describe the

general transient response of the bridge-wire temperature; however, they have not been evaluated for times of several orders of magnitudes lower than the EED’s thermal constant.

In this chapter, the transient thermal response of a typical EED predicted by analytical models is compared with numerical results obtained using ANSYS® Multiphysics. The EED is analyzed considering constant input current. Special emphasis is devoted to time periods much smaller than the thermal constant of the EED. Moreover, the thermal models are used to assess the susceptibility of EEDs with connection wires to different electromagnetic pulsed environments.

### 3.2 EED’s Transient Thermal Models

When an EED is fed, the energy conservation law in the bridge-wire indicates that

\[
\frac{dQ_{\text{gen}}}{dt} - \frac{dQ_{\text{out}}}{dt} = \rho V C_p \frac{d\theta}{dt}
\]  

(3.1)

where \( \frac{dQ_{\text{gen}}}{dt} \) and \( \frac{dQ_{\text{out}}}{dt} \) are the generated and the output heat flow in the bridge-wire, \( \rho \) is its density, \( V \) is its volume, \( C_p \) is its specific heat, and \( \theta \) is its temperature. The generated heat flow corresponds to the dissipated energy due to the current flow in the bridge-wire. The output power represents the heat flow towards the exterior of the bridge-wire. This heat is conducted to the exterior of the EED by means of the materials between the bridge-wire and the casing.

#### 3.2.1 First-order Model

The EED thermal response can be modeled using a first-order differential equation based on experimental exponential cooling curves [11].

\[
p(t) - \gamma \theta = C \frac{d\theta}{dt}
\]

(3.2)

In (3.2), the output heat flow \( \frac{dQ_{\text{out}}}{dt} \) is related to the temperature in the bridge-wire by an equivalent thermal conductance \( \gamma \), which represents the material capability to conduct heat between the bridge-wire and the exterior of the EED. In the same way, the coefficient of the time derivative in (3.1) is replaced by the thermal capacitance \( C \).

Solving (3.2) for a constant power \( p(t) = P_0 u(t) \), where \( u(t) \) is the unit-step function, the resulting temperature increase is given by

\[
\theta(t) = \frac{P_0}{\gamma} \left(1 - e^{-t/\tau}\right)
\]

(3.3)

where \( \tau = C/\gamma \) is the thermal constant of the EED.
Table 3.1 presents values of $\gamma$, $C$ and $\tau$ of EEDs measured by different authors. In almost all the cases, the characterized EEDs are commercially available. The only exception is the case reported in [34], where the measured device is an EED without explosives and filled with an inert material. The EEDs in Table 3.1 were classified to low energy and high energy depending on the value of the capacitance.

<table>
<thead>
<tr>
<th>Type</th>
<th>Reference</th>
<th>Heat losses $\gamma$ [mW/°C]</th>
<th>Capacitance $C$ [µJ/°C]</th>
<th>$\tau$ [ms]</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>Average</td>
<td>SD</td>
<td>Average</td>
</tr>
<tr>
<td>Low</td>
<td>[43]</td>
<td>0.732</td>
<td>0.101</td>
<td>39.1</td>
</tr>
<tr>
<td>Low</td>
<td>[44] Type 1*</td>
<td>1.937</td>
<td>1.077</td>
<td>44.5</td>
</tr>
<tr>
<td>High</td>
<td>[44] Type 2*</td>
<td>3.914</td>
<td>0.594</td>
<td>165.6</td>
</tr>
<tr>
<td>High</td>
<td>[10]</td>
<td>0.045</td>
<td>4e-4</td>
<td>300</td>
</tr>
<tr>
<td>Low</td>
<td>[34]</td>
<td>0.045</td>
<td>4e-4</td>
<td>300</td>
</tr>
<tr>
<td>Low</td>
<td>[45]**</td>
<td>0.951</td>
<td>31.66</td>
<td>33.3</td>
</tr>
</tbody>
</table>

*Two types of EEDs with different all-fire currents were tested in [44]. Electrical resistivity of Type 1 and Type 2 were $R_1 \approx 2\Omega$ and $R_2 \approx 1.03\Omega$, respectively.

**Parameters calculated from curves presented in [45]

Different techniques were employed for the characterization of the $\gamma$ and $C$ parameters presented in Table 3.1. In [44], the step response of the bridge-wire temperature was measured from the thermal dependence of the electrical resistance of the bridge-wire. The experimental setup was based on a Wheatstone bridge. The values of the thermal constant and heat losses were determined as

$$\tau = \frac{t_{1/2}}{\ln 2} \quad (3.4)$$

$$\gamma = \frac{P_{in}}{\theta_{max}} \quad (3.5)$$

where $t_{1/2}$ is the time up to one half of the maximum temperature $\theta_{max}/2$ from $t=0$. The thermal capacitance can be obtained as $C = \gamma \tau$. The maximum temperature $\theta_{max}$ should be taken as the stable temperature of the exponential curve; however, the best choice of $\theta_{max}$ is different since the experimental curve is not a pure exponential. Applying a least squares fitting, the optimal $\theta_{max}$ is not obtained as the temperature at the end of the exponential but near the middle of the measurement’s temporal span.

A destructive procedure is proposed by Adam et al. in [43]. Assuming that the critical temperature of the primary explosive (lead azide) $\theta_c$ and the ambient temperature $\theta_{oA}$ are known,
the thermal conductance is obtained from the firing power $P_f$. Applying a slow current ramp, the transient factor in (3.2) can be neglected and $\gamma$ can be calculated as

$$\gamma = P_f / (\theta_c - \theta_\infty).$$  \hfill (3.6)

On the other hand, the thermal capacity is obtained with a short pulse (pulse width $= w \ll \tau$).

$$C = U_f / (\theta_c - \theta_\infty),$$  \hfill (3.7)

where $U_f = Pw$ is the energy used to achieve the critical temperature.

In [11, 34], a frequency-domain technique is applied. The EED is used as the load of a bridge-circuit excited with a sinusoidal source. The modulation of the bridge-wire resistance is related to the amplitude of a bridge-circuit voltage. The thermal constant is calculated from the frequency at which the resistance variation does not follow the sinusoidal input voltage.

### 3.2.2 Second-order Model

The first-order approach (Eq. (3.2)) describes the general behavior of the temperature in the bridge-wire. However, some differences with experimental results are observed since the model neglects the temperature increase in the surroundings of the bridge-wire [42]. To include this effect, Prinse and Leeuw proposed a second-order model in [42], in which the increase of the average temperature of the pyrotechnic mixture is incorporated.

$$p(t) = \gamma_1 (\theta_1 - \theta_2) + C_1 \frac{d\theta_1}{dt}$$  \hfill (3.8)

$$\gamma_1 (\theta_1 - \theta_2) = \gamma_2 (\theta_2 - \theta_0) + C_2 \frac{d\theta_2}{dt}$$  \hfill (3.9)

Eqs. (3.8) and (3.9) represent the heat flow from the bridge-wire to the surrounding materials and from the surroundings to the ambient, respectively. In this model, $\theta_0$, $\theta_1$, and $\theta_2$ are the ambient, bridge-wire, and surrounding materials average temperatures, respectively. $\gamma_1$ and $C_1$ are the thermal parameters of the bridge-wire, and $\gamma_2$ and $C_2$ are the parameters of the pyrotechnic mixture. As a consequence, two thermal constants can be calculated: $\tau_1 = C_1 / \gamma_1$ and $\tau_2 = C_2 / \gamma_2$.

Solving (3.8) and (3.9) for a constant input power and assuming $\theta_0 = 0^\circ C$, the bridge-wire temperature yields

$$\theta_1 = P_0 [A_1 (1 - e^{\alpha_1 t}) + A_2 (1 - e^{\alpha_2 t})],$$  \hfill (3.10)

where
3.3 Short Excitation

When a short excitation \((t \ll \tau)\) is directly injected in the bridge-wire, simplifications of the models can be applied. In this case, the heat flow outward the bridge-wire can be neglected; then, both (3.2) and (3.8) can be reduced to

\[
A_1 = \frac{\alpha_1 C_2 + \gamma_1 + \gamma_2}{\alpha_1 (\alpha_2 - \alpha_1)(C_1 C_2)}
\]

\[
A_2 = \frac{\alpha_2 C_2 + \gamma_1 + \gamma_2}{\alpha_2 (\alpha_1 - \alpha_2)(C_1 C_2)}
\]

\[
\alpha_{1,2} = \frac{1}{2} \left( -b \mp \sqrt{b^2 - 4 \frac{\gamma_1 \gamma_2}{C_1 C_2}} \right)
\]

\[
b = \frac{C_1 \gamma_1 + C_1 \gamma_2 + C_2 \gamma_1}{C_1 C_2}
\]

Thus, the final temperature of the bridge-wire can be predicted assuming a constant excitation with the equivalent average amplitude. This is a consequence of the simplification of the process to an adiabatic one in the short excitation condition.

\[
p(t) \approx C_T \frac{d\theta}{dt}
\]

(3.11)

where \(C_T\) is equal to \(C\) for the first order model and \(C_1\) for the second-order model. Then, the bridge-wire temperature is given by

\[
\theta = \frac{1}{C_T} \int p(t) dt
\]

(3.12)

Representing the excitation power by the equivalent average power \(P_{ave}\), the temperature at the time \(t\) is given by

\[
\theta = \frac{1}{C_T} P_{ave} t
\]

(3.13)

3.4 Repetition Gain

Most of high-power sources deliver their energy in repetitive pulses. Each coupled pulse in the EED generates an increase in the bridge-wire temperature. However, since the excitation is not continuous, the bridge-wire presents repetitive heating and cooling cycles. An additive effect of the pulses in the temperature can be obtained only if the period of the pulses \(T\) (the inverse of the pulse repetition frequency (PRF)) is smaller than about three times the EED’s thermal constants \(\tau\).
To illustrate this effect, consider a train of pulses of constant power $P_0$ (a non-constant signal could also be considered with the same procedure by replacing $P_0$ by its average power during the “on” period). Using the first-order thermal model, the temperature increase due to the first pulse is given by

$$\theta(\tau_r) = \frac{P_0}{V}(1 - e^{-\tau_r/\tau}) \quad (3.14)$$

where $\tau_r$ is the pulse duration. After the excitation, during the “off” period $t_{off}$, the bridge-wire cools down to the temperature given by:

$$\theta(\tau_r + t_{off}) = \theta_1 e^{-(T-\tau_r)/\tau} \quad (3.15)$$

This corresponds to the initial temperature at the bridge-wire before the second pulse excitation. By considering this initial condition, the temperature increase due to the second pulse, and the cooling process, the temperature after the second pulse can be readily calculated as:

$$\theta(2T - t_{off}) = \theta(\tau_r)(1 + e^{-T/\tau})$$

Following the same procedure, the temperature after the $k^{th}$ pulse yields

$$\theta(kT - t_{off}) = \theta(\tau_r) \left(1 + e^{T/\tau} + \ldots + e^{-(T/\tau)(k-1)}\right)$$

or

$$\theta(kT - t_{off}) = \theta(\tau_r) \frac{1 - e^{-(T/\tau)k}}{1 - e^{-T/\tau}} \quad (3.16)$$

From (3.16), it is possible to define

$$G_R(k) = \frac{1 - e^{-(T/\tau)k}}{1 - e^{-T/\tau}}$$

which represents the temperature gain produced by $k$ repetitive pulses as compared with a single pulse. The repetition gain is presented in Fig. 3.1 for different $T/\tau$ ratios. This result agrees with the response of the energy accumulation due to a train of rectangular pulses presented by Netzer in [40]. When the number of pulses goes to infinity, the gain converges to

$$G_{RT} = G_R(k \to \infty) \approx \frac{1}{(1 - e^{-T/\tau})}$$

This factor, here denoted as the total repetition gain, quantifies the difference of having a single pulse source and a repetitive source. When the pulse period $T$ is greater than three thermal constants, the repetition gain tends to unity and, as a consequence, the temperature increase due to a pulse train is approximately the same as that produced by a single pulse.
To compare the temperature curves predicted by the first-order and second-order models, a 3D model of a typical EED was implemented in ANSYS. The dimensions and properties of the simulated EED were based on the state-of-the-art blasting cap (SABC) model [13] and are presented in Fig. 2.1 and Table 3.2, respectively. The simulation included both natural-convective heat transfer with the surrounding air and conductive heat transfer inside the EED.

### Table 3.2 Thermal Characteristics of the Modeled EED

<table>
<thead>
<tr>
<th>Material</th>
<th>Density [kg/m³]</th>
<th>Specific heat [J/kg*K]</th>
<th>Conductivity [W/m*K]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Platinum iridium alloy*</td>
<td>21700</td>
<td>485</td>
<td>14</td>
</tr>
<tr>
<td>Rubber</td>
<td>1000</td>
<td>1800 [46]</td>
<td>0.6 [10]</td>
</tr>
<tr>
<td>Lead azide</td>
<td>4710</td>
<td>569.02 [47]</td>
<td>0.176 [10]</td>
</tr>
<tr>
<td>PETN</td>
<td>1770</td>
<td>1138 [48]</td>
<td>0.17 [10]</td>
</tr>
</tbody>
</table>

*The alloy Pt10Ir [49] was used to represent the bridge-wire.

In order to include the natural convection of the air in the simulation, the convection coefficient was calculated by simulating the structure and the surrounding air in steady state with the tool FLOTTRAN. Fluid elements FLUID142 were used to mesh the structure, allowing the simulation of the fluid behavior of the air. A mean convection coefficient of 7.16 W/m²°C was obtained over the EED’s metallic surface. Then, this coefficient was used to include the effect of the convective heat flux; as a result, the simulation was reduced only to the conductive process inside the EED. In this case, 3D solid elements SOLID87 were used for the meshing.

The structure was excited applying a power generation condition in the bridge-wire. The meshing size was reduced to comply the following condition [50]:

![Fig. 3.1 Repetition gain as a function of the number of pulses](image-url)
where $\Delta l$ is the conducting length in the highest temperature gradient and $k$ is the thermal conductivity. Since the early-time responses due to excitations with different waveforms were of interest, this restriction resulted in a high density of elements near the bridge-wire.

### 3.6 Analytical Model Validation

Different types of signals were used as input power to validate the analytical results. For the comparison, we used both the first-order and the second-order analytical models, where the parameters were obtained using the procedures explained in Section 3.2, and summarized hereunder:

i. Second-order model where the parameters (i.e. $y_1, y_2, C_1, \text{ and } C_2$) are obtained from a least-square fit of the step response of the ANSYS model.

ii. First-order model where the parameters (i.e. $\gamma$ and $C$) are obtained from a least-square fit of the step response of the ANSYS model.

iii. First-order model following the procedure proposed by Kankane et al. in [44]. The thermal parameters are calculated from the step response of the ANSYS model.

iv. First-order model following the procedure proposed by Adam et al. in [43]. A pulse with a width of 1ms is applied to the ANSYS model to calculate the thermal capacity. The thermal conductance is calculated by using a slow power ramp.

Fig. 3.2 a) Late-time and b) early-time responses of the bridge-wire temperature increase due to a constant input power of 0.1W.
Table 3.3 Calculated parameters for the First-order models

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>$\tau$ [ms]</td>
<td>94.1</td>
<td>57.7</td>
<td>18.1</td>
</tr>
</tbody>
</table>

Table 3.4 Calculated parameters for the Second-order models

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Least Squares Fit</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\tau_1$ [ms]</td>
<td>19.3</td>
</tr>
<tr>
<td>$\gamma_1$ [mW/°C]</td>
<td>1.5</td>
</tr>
<tr>
<td>$C_1$ [$\mu$J/°C]</td>
<td>28.91</td>
</tr>
<tr>
<td>$\tau_2$ [ms]</td>
<td>190</td>
</tr>
<tr>
<td>$\gamma_2$ [mW/°C]</td>
<td>2.3</td>
</tr>
<tr>
<td>$C_2$ [$\mu$J/°C]</td>
<td>437.1</td>
</tr>
</tbody>
</table>

The parameters calculated with the different approaches are reported in Table 3.3 and Table 3.4. These values are within the range of the measured data for low energy EEDs presented in Table 3.1.

A common waveform used for the characterization, as well as for the activation of EEDs is a unit-step function. The model in ANSYS was used to calculate the expected response. These results were verified with measurements presented in [43] and [44].

In Fig. 3.1a, the results obtained using ANSYS are compared with the different approaches of first and second order. The simulated window was enough to cover five time constants $5\tau_2$ of the slowest exponential of the second order model, as indicated in the figure. Evidently, the second order model presents the best overall fit to the simulated response. However, the best fit in the early-time response (i.e. $t \ll \tau$) is obtained with the first-order model based on the procedure presented in [43], as shown in Fig. 3.2b. This is because, in this approach, the thermal capacitance and conductance used are independently calculated. Thus, following the procedure proposed in [43] assures that the capacitance really represents the inverse of the slope of the temperature as a function of the time for early times ($t \ll \tau_1$). On the contrary, the capacitance in the other three methods also becomes dependent on the late response.

In fact, the capacitance of this approach is the closest to the theoretical value obtained from (1)

$$C_{\text{theo.}} = \rho V C_p = 15.5 \, \mu J/\degree C$$

where $\rho$, $V$, and $C_p$ are the properties of the bridge-wire material presented in Table 3.2.
3.7 Electromagnetic Coupling Model

In the previous example, it was assumed that the excitation current was directly injected in the bridge-wire. In this section, a radiated excitation is considered. First, the EED’s detonation condition is presented and, then, the electromagnetic model of an EED with connection wires is described.

3.7.1 EED Ignition

The detonation of an EED is obtained when the primary explosive, which is in contact with the bridge-wire, reaches its auto-ignition temperature. This critical temperature for the lead azide is about $\theta_{\text{crit}} = 350^\circ \text{C}$ [51]. According to the approximation (3.13) for short excitations, the necessary energy to obtain this temperature is

$$U_{\text{crit-short}} = \theta_{\text{crit}} C_T = 6.3 \text{mJ}$$

Therefore, if the energy supplied to the bridge-wire is higher than this value, a detonation will be achieved.

When the duration of the excitation is comparable to $\tau$, an energy comparison is not valid. For this case, it is necessary to calculate the bridge-wire temperature with a complete model and to compare this value with $\theta_{\text{crit}}$.

3.7.2 Electromagnetic Model

In order to consider an incident electromagnetic field as excitation, a transfer function between the incident field and the dissipated power in the bridge-wire must be calculated.

First, the transfer function between the incident electric field $E_{\text{inc}}$ and the induced current in the bridge-wire $I_{bw}$ is calculated in the frequency domain using the equivalent Norton circuit described in Section 2.2. Then, the current in the bridge-wire can be calculated by propagating the input current on the EED to the bridge-wire.

For instance, suppose that an electromagnetic field is impinging an EED with connection wires of 20 cm in dipole configuration. For the sake of simplicity, assume a well polarized plane wave as excitation. The obtained transfer function $I_{bw}/E_{\text{inc}}$, using the high frequency EM coupling model, is shown in Fig. 3.3. The local maxima of the transfer function correspond to the frequencies at which the transmission of energy is the highest. The global maximum for this configuration, which represents the best power coupling, is obtained at 535.5MHz.
Fig. 3.3 The transfer function between an incident electric field and the induced bridge-wire current.

\[ I_{bw}(j\omega) = \left( \frac{I_{bw}}{E_{inc}} \right) E_{exc}(j\omega). \]  
\[ (3.17) \]

Finally, the dissipated power in the bridge-wire electrical resistance, \( R_{bw} \), can be obtained as

\[ p_d(t) = [i_s(t)]^2 R_{bw} \]  
\[ (3.18) \]

where \( i_s(t) \) is the inverse Fourier transform of \( I_s(j\omega) \).

### 3.8 Assessment Against IEMI

In this section, the electromagnetic model and the analytical thermal models are used to assess High Power Electromagnetic (HPEM) environments that could cause the activation of the EED with connection wires. Different kinds of excitations are considered. The signal waveforms and
intensities are based on implemented sources that could be used as Intentional Electromagnetic Interferences (IEMI) [52]. In addition, the considered sources are restricted to be transportable in vehicles. Thus, the used excitations are signals produced by a mesoband radiator, two types of radars, and a hyperband radiator. The spectra of these signals are presented in Fig. 3.4.

Remember that according to the comparisons of the analytical thermal models with the ANSYS simulations, the first-order model describes well the thermal dissipation in the EED for short excitations. On the other hand, for time periods comparable to the EED’s thermal constant, the second order model presents the best response. For this reason, different models are used depending to the duration of the excitation. In the following results, the used parameters of the first-order model are the calculated ones with the Adam et al. procedure [43]. In addition, the time scales of the figures are normalized to the thermal constant of this first-order model.

3.8.1 Mesoband Radiator

A damped sinusoidal excitation tuned in the frequency of the best coupling (535.5MHz) is chosen as incident electric field. The intensity of the incident field is set to 100kV/m and its shape is represented using a double exponential function

\[ E_{ds}(t) = E_0 (e^{-a_1 t} - e^{-a_2 t}) \sin \omega_0 t \]

with attenuation factors \( a_1 = 0.083/\omega_0 \) and \( a_2 = 0.047/\omega_0 \).

These parameters are based on signals produced by a mesoband radiator reported in [53] at distance of 2 m.

The incident electric field, the induced current and the dissipated power in the bridge-wire are presented in Fig. 3.5. Integrating the obtained power, the induced energy in the bridge-wire can be calculated. It reaches a value of

\[ U_{MR} = 1.0 \text{mJ} \]

which, applying the short excitation approximation (8) with the thermal capacity of the first order model, is equivalent to

\[ \theta_{MR} = 55.9^\circ C \]

The obtained energy is below to the activation threshold calculated earlier. In fact, for this case, an electric field magnitude higher than 250kV/m is necessary to reach the critical energy.
3.8.2 Radar

Other scenario that should be also considered is an excitation with pulsed RF emissions, such as the one produced by radars. Although the contribution of a single pulse could not be enough to achieve the critical threshold, the accumulated effect of a series of pulses, with the adequate pulse repetition frequency (PRF), could produce the activation.

Two types of UHF radars are considered. First, the characteristics of a long-pulse radar, which are based on parameters of state-of-the-art solid-state elements used for radar applications [54], are used to calculate the excitation. The system is supposed to be tuned at the EED optimal frequency:
Fig. 3.6 Average temperature responses of the EED’s bridge-wire excited by two radar signals and a series of hyperband pulses.

- Peak output power: 700W
- PRF: 1kHz
- Pulse width: 0.2ms
- Duty cycle: 20%
- Operating frequency: 535.5MHz
- Channel bandwidth: 5MHz
- Antenna Gain: 10 dB

Then, the energy coupled in an EED due to a higher power radar is calculated. The radar AN/SPS-40B [55], normally used in army ships, is considered in this case. Its characteristics are:

- Peak output power: 200kW
- PRF: 300Hz
- Pulse width: 60ms
- Duty cycle: 1.82%
- Operating frequency: 450MHz
- Channel bandwidth: 5MHz
- Antenna Gain: 21 dB

The incident field for each case can be calculated by using

\[ |E_{exc}| = \frac{\sqrt{60P_t G_t}}{r} \]  

(3.19)

where \( P_t \) is the transmitted power, \( G_t \) is the antenna gain, and \( r \) is the distance between the transmitter and the EED.

The dissipated power due a single pulse can be calculated with the coupling model presented before. Then, the average power of a series of pulses can be obtained by multiplying the energy
contribution of one pulse by the PRF. This power value is then inserted in the second order model since the temperature increases with a slow rate.

In Fig. 3.6, the bridge-wire temperature increase due to the incidence of both radar signals is presented. The figure shows that the critical temperature can be achieved with both devices. For the solid-state based radar case, it is necessary to have a distance lower than 2.4 m to achieve the detonation. On the other hand, with the long range radar the threshold is easily reached, even at a distance of 5 m. Notice that the carrier frequency in the second case does not match with the optimal frequency of the EED transfer function, Fig. 3.3 and Fig. 3.4.

### 3.8.3 Hyperband radiator

A hyperband radiator can produce radiated pulses of high power and of very short duration, typically 100 ps or so. As a consequence, the spectrum of this signal covers a wide frequency range, including the frequencies in which the EED’s transfer function is the maximum.

The generation of high power pulses has been studied in the past decade. As a consequence, different devices were conceived; however, the implemented radiator of impulse like waveforms with the highest field-range product ($r_{E_{far}}$) was the JOLT [56]:

- Source Voltage: 1MV
- PRF: 600Hz
- Pulse rise time (radiated field): 80ps
- Full-width half maximum (FWHM): 100ps
- Operating frequency: 50MHz-2GHz
- Field-range product ($r_{E_{far}}$): ~5.3MV

The near field produced by the JOLT system at 5m was calculated following the procedure presented in [57]. This field is inserted in the EM coupling model to calculate the dissipated power induced by a single pulse. Then, the temperature increase can be calculated by using the simplification of the first-order model, Eq. (3.12). These parameters are presented in Fig. 3.7.

The temperature achieved with a single pulse induces a temperature increase of 7.8°C. This result agrees with experimental results that show that hot-wire EEDs have a hyperbolic response of the dissipated power due to pulse excitations as a function of the pulse width [10]. When the pulse width becomes smaller, the firing power tends to infinity.

Since the JOLT system produces a repetitive pulse, the late temperature increase was obtained by calculating the average power, as in the radar case. Fig. 3.6 presents the bridge-wire temperature in the order of several EED’s time constants. Although the incident electric field is very intense, the results show that the maximum achieved temperature is lower than the critical threshold.
The final temperature can be also calculated using the JOLT’s total repetition gain, \( G_{RT} = 11.4 \), and the temperature produced by a single pulse, \( \theta(\tau_r) = 7.8^\circ C \),

\[
\theta_f = G_R \theta(\tau_r) = 88.92^\circ C
\]

A slight underestimation is obtained as compared with the temperature obtained with the average power procedure, \( \theta_f = 89.37^\circ C \), which is shown in Fig. 3.6. This is due to the fact that this approach is based on the first order thermal model, which has not into account the heating of the pyrotechnic mixture.

Fig. 3.7 a) Incident electric field produced by the JOLT radiator and the corresponding bridge-wire induced b) current, c) power, d) and temperature.
A summary of the analyzed sources of IEMI and their effect on an EED is presented in Table 3.5 IEMI Effects on EEDs, in which the main parameters that determine the performance of the transmission of energy are reported. As can be seen, the most disturbing signal corresponds to the one produced by a radar. The EED’s critical temperature was only achieved with the radar signals. Although the intensities of the radiated fields by the considered radars are of several orders lower than those produced by the other sources, the radar signals deliver more total energy to the bridge-wire in the same integration time. One of the main differences is the considerable higher duty cycle.

All the sources, except the mesoband radiator, present a repetition gain greater than 1. The highest gain is presented with the long-pulse radar given to its PRF of 1 kHz. In addition, this source presents the highest duty-cycle and a flat spectrum around the frequency of the EED’s optimal coupling. However, the characteristic of this source is the low output power.

Two different distances (i.e. 2 and 5m) between the radiator and the victim were considered due to the differences in energy consumption and size of the sources. For practical applications (e.g. IED demining), a distance of 5m is the minimum distance for safety reasons; thus, the AN/SPS-40B radar appears to be the most appropriate source.
3.9 Conclusions

The transient thermal response of an EED was calculated using simulations of ANSYS and analytical models. Different analytical approaches reported in the literature to calculate the thermal parameters of first and second order models were assessed taking as reference ANSYS numerical simulations. For the early-time response \((t<<\tau)\), the best approach corresponds to a first-order differential model in which the thermal capacitance is calculated with short-pulse excitations. It was shown that this model adequately describes the bridge-wire temperature for different transient input power signals. A linear simplification to calculate the maximum temperature due to short excitations was also shown to be adequate. On the other hand, the most appropriate model for the late-time response is a second-order model.

The analytical thermal models validated in this work relate, in a simple way, the thermal and the electromagnetic processes. As a result, key EED’s firing conditions are easily evaluated. For short excitations, an energy threshold can be used; while, for long duration excitations, the firing condition is to exceed a threshold in the bridge-wire temperature.

The models were used to assess the electromagnetic susceptibility of an EED with connection wires of 20 cm length in dipole configuration for different HPEM environments. Radiated signals produced by a mesoband radiator, two types of radars, and a hyperband radiator were considered. The obtained results suggest that the radar signal is the most disturbing to achieve the EED detonation because of its higher duty-cycle as compared with the other sources. The temperature firing threshold was exceeded with the radiated field produced by a radar of 200kW of output power located at a distance of 5 meters.
Chapter 4

Susceptibility of Electro-Explosive Devices to Microwave Interference*

In the previous chapter, it was shown that continuous wave (CW) radiated pulsed signals can considerably disturb Electro-Explosive Devices (EEDs) with specific wiring conditions. In this chapter, the electromagnetic susceptibility against CW signals of hot-wire EEDs with arbitrary connection wires is assessed statistically. The electromagnetic coupling and the thermal power dissipation are modeled to determine the activation condition due to an excitation with an external electromagnetic field. The reception properties of the connection wires are obtained numerically and validated experimentally; variations in their geometry are considered by means of a Monte Carlo approach. The optimal coupling frequency and the probability of activation of a typical wired EED as function of the magnitude of the excitation are obtained. A detonation probability of 95% is obtained for a wired EED illuminated with a 2447 V/m incident field.

4.1 Introduction

The activation of EEDs due to EM field exposure depends on the intensity and the spectral content of the incident field. Narrow band systems, used to produce Intentional Electromagnetic Interferences (IEMI), can be efficiently couple high currents in EEDs due to the resonant behavior of EED [58]. If the transfer function between the incident field and the dissipated power in the EED is known, optimal coupling frequencies could be derived. Mora et al. proposed a methodology to calculate the transfer function including the transient energy conversion into heat inside an EED [25]. It can be applied if the complete system characteristics are known. However, actual circuits with EEDs present connection wires with arbitrary geometries and, as a consequence, arbitrary frequency responses. This randomness in the system calls for a statistical analysis of the problem.

Monte-Carlo approach has been a useful tool to characterize electromagnetic interactions with wires. The response of twisted-wire pairs exited by a plane wave was studied by Armenta et al. [59], where the effect of small random variations of the twisting was analyzed through this method.

in a wide frequency range. Faster techniques that take advantage of the statistical characteristics of the problem, but with the same principle, have also been proposed. For example, the statistical indicators (i.e. mean, standard deviation and kurtosis) of the induced voltage in an undulating thin-wire over a ground plane have been obtained with a similar method denominated sparse grid [60].

Other approaches, based in the computation of an explicit relationship between the observable and the random characteristics, have been proposed recently. These are called probabilistic approaches; on the contrary, Monte-Carlo method corresponds to a statistical analysis [61]. Some of them, presented in depth in [62], include perturbation method, polynomial chaos, semi-intrusive method, the use of maximum entropy, and Discrete-Inverse-Fourier-Transform (DIFT). They are applicable in problems where the observable moments are not calculable analytically, such in the analysis of radiated susceptibility. The objective of these approaches is to use statistical strategies to reduce the number of samples (simulations) in the deduction of the statistical moments of observables.

In this chapter, the probability of activation of an EED with connection wires with arbitrary geometry and excited by a linearly polarized electromagnetic field is presented. First, the electromagnetic (EM) coupling of a plane wave on an EED with wires is analyzed deterministically. In the third section, a steady state thermal model, based on ANSYS® simulations, to find the activation dissipated power is depicted. In the fourth section, a statistical analysis of the induced power in an EED illuminated with a continuous-wave electric-field and with random configurations of wires is developed with a Monte-Carlo approach. The frequency dependence of the mean induced power and the survivor function of the samples as a function of the field magnitude are obtained. Finally, a discussion on the obtained results and general conclusions are presented in section five.

4.2 Electromagnetic Response

The coupling between a source of EM disturbance and a receiver, namely victim, is determined by the efficiency of the transfer path. This path can be represented by a series of transfer functions in cascade, allowing the analysis of the problem by blocks [6]:

\[ \text{Source} \rightarrow \text{Antenna} \rightarrow \text{Propagation} \rightarrow \text{External Interaction} \rightarrow \text{Penetration} \rightarrow \text{Port of Interest} \]

Therefore, the EM coupling between an incident plane wave and the bridge-wire of an EED can be decomposed in two transfer functions [25, 63], which correspond to the external interaction and the penetration. The port of interest, in the case of EEDs, corresponds to the conversion of electric current into heat given in the bridge-wire.
4.2.1 External Interaction: Plane Wave Excitation

The coupling of an EM wave impinging the EED can be calculated using the Norton equivalent presented in Section 2.2. Although the Norton source and the source equivalent impedance could be calculated by using analytical expressions for canonical cases (e.g. dipole), the parameters of a wire with arbitrary geometry must be obtained through a full-wave simulation. The Time Domain Integral Equation (TDIE) technique with Marching on Time (MOT) scheme [64] was chosen in this study. The antenna input impedance can be calculated exciting the structure with a voltage source located on the mid-point of the structure, and the induced short circuit current (i.e. Norton current) can be determined by exciting the complete geometry of the wires with an incident plane wave. The input impedance of the EED was calculated analytically by means of the transmission line model [13] explained in Chapter 2.

4.2.2 Penetration

The actual element in the system that transforms EM energy into heat in the EED is the bridge-wire (see Fig. 2.1). Thus, the activation state depends on the dissipated power on this element. Neglecting the losses in the transmission lines that represent the EED, the power dissipated on the bridge-wire is the same as the one delivered to the input of the EED. Therefore, the dissipated power in the EED \( P_d \) can be obtained using (2.11). In addition, considering the wires as an antenna, this power can be calculated as [21]

\[
P_d = W_i A_{er},
\]

where \( W_i = |E_i|^2/(240\pi) \) is the power density of the incident electromagnetic wave, \( E_i \) is the incident electric field, and \( A_{er} \) is the realized effective area, which only depends on the antenna and load properties. Note that the impedance mismatch and the polarization mismatch losses are included in \( A_{er} \). Thus, the realized effective area corresponds to the transfer function of the electromagnetic coupling between the incident plane wave and the bridge-wire power. \( A_{er} \) can be calculated from the power obtained with (2.11) and the known incident electric field.

4.3 Point of Interest: Steady-State Thermal Dissipation

EEDs, specifically hot-wire detonators, are activated by a deflagration process [65]. When the temperature of the bridge-wire exceeds the autoignition temperature of the primary charge, it detonates and activates the secondary explosive. This critical temperature for the lead azide, a material commonly used as primary charge, is 350 °C [51]. By using a 3D thermal model
implemented in ANSYS®, the necessary power dissipated in the bridge-wire to achieve the critical temperature was calculated.

The steady state model includes both natural convective heat transfer with the surrounding air and conductive heat transfer inside the EED. In order to include the natural convection of the air in the simulation, the structure and the surrounding air were implemented in ANSYS® Mechanical APDL application with FLUID142 type elements and the solution was obtained using a FLOTRAN analysis. The dimensions of the simulated EED are given in Fig. 2.1 and the properties of the implemented materials in the thermal model are depicted in Table 4.1. The model was excited applying a constant power generation condition in the bridge-wire.

<table>
<thead>
<tr>
<th>Material</th>
<th>Density (kg/m$^3$)</th>
<th>Specific heat (J/kg*K)</th>
<th>Conductivity (W/m*K)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Air</td>
<td>AIR-SI$^a$</td>
<td>AIR-SI$^a$</td>
<td>AIR-SI$^a$</td>
</tr>
<tr>
<td>Aluminum</td>
<td>2702</td>
<td>903</td>
<td>237</td>
</tr>
<tr>
<td>Rubber</td>
<td>1000</td>
<td>1800$^{22}$</td>
<td>0.6$^3$</td>
</tr>
<tr>
<td>Lead azide</td>
<td>4710</td>
<td>569.02$^{23}$</td>
<td>0.176$^3$</td>
</tr>
<tr>
<td>PETN</td>
<td>1770</td>
<td>1138$^{24}$</td>
<td>0.17$^5$</td>
</tr>
</tbody>
</table>

$^a$ Air material in the ANSYS® FLOTRAN analysis

A power sweep was performed in the simulation to obtain the bridge-wire temperature as a function of the dissipated power. The simulated temperature increment, presented in Fig. 4.1, shows a linear dependence with the dissipated power with a slope of 973.76 °C/W. Therefore, the temperature increase $\Delta \theta$ can be expressed as

$$\Delta \theta = R_{th} P_d$$  \hspace{1cm} (4.2)

where $P_d$ is the dissipated power and $R_{th}$ is the EED’s thermal resistance (i.e. the inverse of the heat loss factor explained in Section 3.2), which corresponds to the slope value of the curve in Fig. 4.1. With this expression, the temperature due to a given power can be directly calculated and vice-versa. Assuming an ambient temperature of 20 °C and $R_{th} = 973.76$ °C/W, the critical temperature of the lead azide is obtained with 0.34 W.

For illustrative purposes, the temperature dependence obtained from the experimental data presented in [44] for a “Type 1” detonator is also plotted in Fig. 4.1. The detonator “Type 1” corresponds to a low energy detonator with low no-fire current (<0.3 mA), such as the one simulated in this work. The measurements were made by feeding the detonator with D.C. power values below 0.05 W and by calculating the bridge-wire temperature increase from its resistance change. The curve presented in Fig. 4.1, with triangular markers, corresponds to the linear extrapolation up to 0.35 W of the measured steady state temperature. Both curves, from ANSYS®
simulation and from experimental results, show similar slopes in the linear responses, which was expected since both are low energy devices.

Fig. 4.1 Bridge-wire temperature as function of the dissipated power. The ANSYS® results are compared with the measurements on a “Type 1” EED [44].

4.4 Statistical Analysis: Wires with Random Geometries

4.4.1 Random Geometries

The variation in the geometries of the connection wires was performed by modifying its length and pattern. Arbitrary patterns were obtained by dividing the wire in sections with equal length but with different inclination angles. Fig. 4.2 shows the six variable angles in the x-y and y-z plane.

Fig. 4.2 Variable angles for the generation of arbitrary geometries of wires

A sample of 500 arbitrary wire structures was simulated with TDIE. These configurations aim to represent typical connections of Improvised Explosive Devices (IEDs) and they were obtained by assigning a random uniformly distributed value to the total wire’s length and to each inclination angle. The variation range of each variable is presented in Table 4.2. The electric field was simulated with direction of propagation $\hat{x}$ and polarization $\hat{y}$. For each case, the short circuit induced current, due to the external incident field, and the wire’s input impedance were calculated. With these values and with the EED’s input impedance obtained with the transmission line model, the power delivered to the bridge-wire of the EED was calculated by using (2.11).
### Table 4.2 Parameters of the EM Coupling Simulation.

<table>
<thead>
<tr>
<th>Variable</th>
<th>Parameter</th>
<th>Min. value</th>
<th>Max. value</th>
</tr>
</thead>
<tbody>
<tr>
<td>L</td>
<td>Total wire’s length</td>
<td>5 cm</td>
<td>25 cm</td>
</tr>
<tr>
<td>$\theta_{1A}, \theta_{1B}$</td>
<td>Inclination angles $\theta_{1A}$ and $\theta_{1B}$</td>
<td>$20^\circ$</td>
<td>$160^\circ$</td>
</tr>
<tr>
<td>$\theta_{2A}, \theta_{2B}$</td>
<td>Inclination angles $\theta_{2A}$ and $\theta_{2B}$</td>
<td>$-90^\circ$</td>
<td>$90^\circ$</td>
</tr>
<tr>
<td>$\phi_A$</td>
<td>Inclination angle $\phi_A$</td>
<td>$20^\circ$</td>
<td>$160^\circ$</td>
</tr>
<tr>
<td>$\phi_B$</td>
<td>Inclination angle $\phi_B$</td>
<td>$-20^\circ$</td>
<td>$-160^\circ$</td>
</tr>
<tr>
<td>a</td>
<td>Wire’s radius</td>
<td>0.7 mm</td>
<td></td>
</tr>
<tr>
<td>$</td>
<td>E_i</td>
<td>$</td>
<td>Incident electric field</td>
</tr>
</tbody>
</table>

#### 4.4.2 Frequency of Optimal Coupling

An optimal frequency of coupling can be defined when the induced power is the maximum possible for most of the observations. The mean value and the confidence interval of a standard deviation of the induced power in the EED as function of the frequency are presented Fig. 4.3a. It clearly shows an optimal frequency range between 500 MHz and 660 MHz in which the mean value of the power increases considerably as compared with the rest of the frequencies.

Although the estimated frequency range limits the spectrum in which a good coupling between the incident electric field and the bridge-wire is probable, its bandwidth is wide for a high-power microwave source. For this reason, the best single frequency must be chosen. In Fig. 4.3b, the probability density function (PDF) of the induced power in the bridge-wire when the incident electric field has a frequency of 600 MHz is presented. This is compared with the best case, which corresponds to having an electric field able to excite the frequency of maximum coupling of each observation, and with an excitation frequency of 2.45 GHz, which is outside the range of good coupling. The figure shows that a normal distribution can represent each case and that the 600 MHz frequency for the excitation increases considerably the induced power in the bridge-wire compared with an arbitrary frequency outside the found range. The mean value of the 600 MHz distribution is 8.9 dB lower than the mean value of the best case, while the difference between the arbitrary case of 2.45 GHz and the best one is 29.2 dB.

#### 4.4.3 Magnitude of Excitation

Now, it is necessary to know the intensity of the field able to induce enough average power to cause a detonation in the observations. To obtain this, the realized effective area of each wire configuration was calculated introducing the simulated power and 1V/m incident field in (4.1). Then, the critical power obtained with ANSYS® (0.34W) was used in the same expression and the magnitude of the electric field necessary to produce the detonation was calculated.
Fig. 4.3 a) Mean value and 68.3% confidence interval of the induced power in the bridge-wire of an EED with connection wires with arbitrary geometries as functions of the frequency. b) Histograms of the induced power for two frequencies: 0.6 GHz and 2.45 GHz. The histogram of the maximum possible induced power is also presented for comparison.

Fig. 4.4 Survivor function of the observations as function of the magnitude of the incident field for a fixed excitation frequency of 0.6 GHz.
In Fig. 4.4, the empirical survivor function of the samples when the incident field is tuned at 600 MHz is shown. This figure depicts the probability of a wired EED to bear a specific electric field intensity without detonating. According to these results, the electric field intensity required to detonate 70% of devices is 890 V/m and to detonate the 95% is 2447 V/m.

4.5 Discussion and Conclusion

A statistical analysis of the susceptibility of EEDs to microwave radiation was presented. Arbitrary geometries in the connection wires of this device were simulated, creating a set of possible configurations. The frequency dependence of the electromagnetic responses of both the connection wires and the EED was included in a coupling model. Then, the transfer function between an incident electric field with linear polarization and the dissipated power in the EED’s bridge-wire was obtained for each sample. Analyzing the probability density of the dissipated power, the optimal coupling frequency of the system for a continuous wave (CW) excitation was calculated. In addition, this power was compared with the critical activation value, which was determined with an ANSYS® steady state thermal model, and the survivor function of a typical EED with connection wires as a function of the magnitude of the incident field was determined.

The separation of the problem in the scheme of external interaction and the penetration provides an easier way to bear with devices that have elements with both random and deterministic (typical) responses. This scheme reduces the complexity of the problem since only the variable part of the problem is simulated numerically and the other is calculated analytically.

Here, a CW excitation was considered; therefore, the activation was determined using a comparison between the average power induced due to the CW excitation and the power threshold obtained in ANSYS® with a steady state thermal model. As discussed in Chapter 3, the analysis of transient excitations (e.g. a damped sinusoidal) requires solving first or second order differential equations and the activation should be determined when the bridge-wire temperature exceeds the critical temperature of the primary charge.

Other factors of the coupling, such as the polarization and the propagation medium, evidently affect the induced power in the EED. The incident wave polarization has a significant effect on the wire’s effective area. In this study, the incident wave was assumed to have the same polarization as the average wire’s geometry. Although, due to the variability of the geometries, there is always a non-zero probability of inducing antenna currents, a considerable reduction in the mean induced power would be expected if an orthogonal polarization were considered. On the other hand, the simulations were carried out in free space; however, different propagation media can be presented.
in a real scenario. In the case of a buried device, for example, three main effects should be considered: i) a reflection of the incident wave due to change of media, ii) attenuation due to the media losses, and iii) disturbance of the device’s near field due to the properties of the surrounding medium. As a result, the induced power in the EED would decrease according to the propagation losses. In addition, the frequency dependence of the mean induced power, as shown in Fig. 4.3a, would result modified caused by the effect of the surrounding media on the wire’s effective length.

In Chapter 6, real samples of firing circuits used in Improvised Explosive Devices are characterized using the technique presented in the next chapter. Effects on the coupling of additional elements in the system, such as the container, are discussed in that analysis.
Chapter 5

Wideband Experimental Characterization of Differential Devices*

A procedure to characterize the gain and the input impedance of differential devices from a mixed S parameters measurement is presented in this chapter. Measurements are performed with a differential probe to connect differential antennas under test and the single-ended ports of a vector network analyzer. This element is modeled with transmission lines in cascade and its effect in the measurement is corrected. Three wire antennas are characterized and a good agreement is found between experimental data and numerical results obtained using the Time Domain Integral Equation technique. The application of this technique to measure the impedance of other differential loads, such an EED, is also shown.

5.1 Introduction

The high-frequency characterization of differential devices, such as antennas or transmission lines, shows some challenges because the measuring instruments usually present unbalanced ports. One alternative to perform this characterization is by using a balun, but it usually has a narrowband behavior [66, 67] and the accuracy of the measurement depends on the accuracy of this device [68]. Another technique is by measuring the S parameters of a differential port conformed by two single-ended ports [69].

The second option provides a wideband measurement and its accuracy depends on the standard calibration kit of the vector network analyzer (VNA) and on the correction procedure due to the insertion of the adapter. This post-processing correction, called de-embedding, removes from the measurement the effect of the jig adapter, which is used to connect the two arms of the differential device, and to provide a virtual ground plane in the middle of the differential port [70]. The adapter affords a transition between two single-ended coaxial ports and a parallel-wire balanced port. It usually consists of a pair of coaxial cables that provide a connection point to the VNA ports and reduces their separation distance to achieve the required one by the antenna input.

De-embedding of the measurement can be performed if the S parameters of the adapter are known. The delay and the attenuation due to the coaxial extensions can be measured from the reflection coefficient when each coaxial cable is terminated in a short circuit. Then, the de-embedding calculation can be performed by post-processing [69] or by using a port extension function of the VNA [66, 68]. In [71], the adapter is terminated using two female SMA connectors, which makes the calibration at the end of the probe possible with a standard SMA kit. However, these techniques do not remove the effect of the pins that connect the coaxial inner conductor with the antenna, which could be considerable in small antennas and wire antennas.

Here, it is proposed replacing the pins of the conventional jig adapter by SMA inner contacts, providing the possibility of performing measurements without soldering the adapter to the device under test (DUT). The adapter is modeled by transmission lines in cascade, whose parameters are obtained experimentally. The proposed adapter and its de-embedding procedure are used to measure the gain and the input impedance of wire antennas and the input impedance of EEDs. The accuracy of the measurements is validated with the simulation of the wire antennas using the Time Domain Integral Equation (TDIE) technique [72].

5.2 Experimental Setup

The gain and the input impedance of differential antennas were characterized using the experimental setup presented in Fig. 5.1. A Rohde & Schwarz ZVB VNA was configured using two logical ports: a balanced port and a single-ended port, as shown in the figure. The received power and the input impedance in the balanced port were calculated from the mixed mode S parameters Ssd21 and Sdd1, respectively.

The measurements were performed in the anechoic chamber of the Universidad de Los Andes, Bogotá. An incident plane wave was generated connecting a reference antenna to the single-ended port of the VNA with a wideband amplifier. Two reference antennas from ETS-Lindgren were used in order to cover the frequency range between 200MHz and 3GHz: a Log Periodic Dipole Array (LPDA) Model 3148B and a Double-Ridged Guide Antenna (HORN) Model 3117.
The antenna under test (AUT) was connected using the adapter depicted in Fig. 5.2. The adapter can be represented by two coaxial segments in parallel connected to a parallel-wire transmission line. The coaxial segments are formed by a SMA f-f adapter, which are held to a dielectric sheet, and two coaxial cables, whose external conductors are soldered in their ends. The parallel transmission line corresponds to the SMA contacts used to connect the AUT without soldering.
5.3 Mixed Mode S-Parameters

The mixed mode S parameters matrix $S_{mm}$, resulting from the measurement using the setup proposed in Fig. 5.3, relates the traveling waves of the single-ended port (logical port 1) and the differential and common mode waves of the balanced port (logical port 2):

$$B_{mm} = S_{mm} A_{mm}$$  \hspace{1cm} (5.1)

In (5.1), $B_{mm} = [b_s, b_d, b_c]'$ and $A_{mm} = [a_s, a_d, a_c]'$ are the response and stimulus waves, respectively, and

$$S_{mm} = \begin{bmatrix} S_{ss} & S_{sd} & S_{sc} \\ S_{ds} & S_{dd} & S_{dc} \\ S_{cs} & S_{cd} & S_{cc} \end{bmatrix}.$$

These mixed mode waves can be related with the single-ended waves in the physical ports by means of the matrix $M$ [73]

$$A_{mm} = MA_{std}$$ \hspace{1cm} (5.2)

or

$$\begin{bmatrix} a_s \\ a_d \\ a_c \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} \sqrt{2} & 0 & 0 \\ 0 & 1 & -1 \\ 0 & 1 & 1 \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \\ a_3 \end{bmatrix}$$

and

$$B_{mm} = MB_{std}$$ \hspace{1cm} (5.3)

or
The transformation between the mixed mode and the standard mode matrices can be obtained by replacing (5.2) and (5.3) in (5.1)

\[ MB_{std} = S_{mm}M_{std} \]

and using the definition of the standard S parameters

\[ MS_{std} = S_{mm}M \]

\[ S_{mm} = MS_{std}M^{-1} \] (5.4)

Since the calibration of the VNA is not performed at the actual port of the differential antenna, it was necessary to apply de-embedding in the balanced port.

### 5.4 Adapter De-embedding

In Fig. 5.3, a block diagram of the elements presented in the measurement is shown. Knowing the S parameters of each block, the T parameters can be calculated and the de-embedding from the calibrated plane to the antenna plane can be performed [74]. The measured T parameters are given by

\[ T_{mm} = T_{adapter}T_{Link} \] (5.5)

where \( T_{mm}, T_{Link}, \) and \( T_{adapter} \) are the measured, de-embedded, and adapter’s T parameters, respectively. Therefore, the corrected \( T_{Link} \) parameters can be obtained by multiplying the measured parameters by the inverse of the mixed T parameters of the adapter

\[ T_{Link} = T_{adapter}^{-1}T_{mm}. \] (5.6)

The T parameters of the adapter correspond to the equivalent T matrix of the coaxial and parallel transmission line blocks presented in Fig. 5.3. The S parameters of a lossy unmatched transmission line with a characteristic impedance \( Z_c \) are given by [75]

\[ S = \frac{1}{D_s} \begin{bmatrix} (Z_c^2 - Z_0^2) \sinh \gamma l & 2Z_0 Z_c \\ 2Z_0 Z_c & (Z_c^2 - Z_0^2) \sinh \gamma l \end{bmatrix} \] (5.7)

where \( D_s = 2Z_0 Z_c \cosh \gamma l + (Z_c^2 + Z_0^2) \sinh \gamma l \), \( Z_0 \) is the characteristic impedance in the measurement system, and \( \gamma \) is the complex propagation constant of the line.
5.4.1 Coaxial Sections

Since the impedance of the used coaxial cables is equal to the system impedance, the S parameters of these blocks depend only on the losses and phase change. They were measured from the reflection coefficient when each coaxial cable was terminated in a short circuit with a soldered conducting plate.

5.4.2 Parallel-wire section

Assuming that the parallel-wire section is lossless and that the propagation velocity is the speed of light, the only unknown in (6) for this line is the characteristic impedance. This was calculated from the measurements of the input impedance when the line was terminated in an open circuit \( Z_{op} \) and in a short circuit \( Z_{sc} \):

\[
Z_c = \sqrt{Z_{op}Z_{sc}}
\]  
(5.8)

where \( Z_{sc} = jZ_c \tan(\beta l) \), \( Z_{op} = -jZ_c \cot(\beta l) \), and \( l \) is the length of the line. These impedances were measured setting a balanced port in the VNA, calibrating with a SMA calibration kit, de-embedding the coaxial sections, and terminating the adapter in open circuit and in short circuit by means of a small rectangular conducting plate. The obtained characteristic impedance is shown in Fig. 5.4, which presents a mean value of 150.78 \( \Omega \) in the real part. It agrees with the theoretical value of 152.57 \( \Omega \), obtained using [76]

\[
Z_c = \frac{1}{\pi} \sqrt{\frac{\mu_0}{\varepsilon_0}} \text{acosh} \frac{s_p}{d_p}
\]

(5.9)

where \( s_p = 2.5 \text{ mm} \) and \( d_p = 1.5 \text{ mm} \) are the separation and the diameter of the pins, respectively.

The obtained characteristics of coaxial and parallel transmission lines are reported in Table 5.1.

![Fig. 5.4. Characteristic impedance of the parallel-wire transmission line formed by the adapter pins.](image-url)
5.4.3 Procedure

The de-embedding of the coaxial blocks was done by using the function “single-ended port de-embedding” of the VNA with the obtained parameters in each case. Then, the data provided by the VNA include the effect of the coaxial extensions and, as a consequence, the matrix $T_{\text{adapter}}$ corresponds to the parallel-wire block. This matrix corresponds to a balanced two-port network.

In summary, the measurement procedure can be summarized as follows:

1. Calibrate the VNA with standard calibration kit at the terminals of the cables.
2. De-embed the coaxial sections with the VNA function “single-ended port de-embedding” and using their previously measured parameters (Table 5.1).
3. Measure the standard S parameters of the 3-port network.
4. Calculate the mixed mode S parameters.
5. Calculate the mixed $T_{\text{mm}}$ parameters of the measurement data.
6. Calculate the corrected T parameters pre-multiplying by $T_{\text{adapter}}^{-1}$, Eq. (5.6).

To obtain good results in the impedance measurement, the cables must be maintained in the same position during the measurement and the calibration.

5.5 Equations

5.5.1 Input Impedance

The differential impedance of the antenna was calculated using (5.10). Here, $Z_{\text{odiff}}$ is the VNA differential impedance and $S_{dd}'$ is the differential reflection parameter of the $S_{\text{Link}}$ matrix, which corresponds to the transformation of the $T_{\text{Link}}$ matrix obtained after the de-embedding procedure.

$$Z_{\text{in}} = Z_{\text{odiff}} \frac{1 + S_{dd}'}{1 - S_{dd}'}$$  \hspace{1cm} (5.10)

5.5.2 Gain

The realized effective area of an antenna is given by

---

Table 5.1 Properties of the transmission lines in the adapter

<table>
<thead>
<tr>
<th></th>
<th>Coaxial 1</th>
<th>Coaxial 2</th>
<th>Parallel Wires</th>
</tr>
</thead>
<tbody>
<tr>
<td>Characteristic Impedance ($\Omega$)</td>
<td>50</td>
<td>50</td>
<td>150.8</td>
</tr>
<tr>
<td>Length / $vf^*$ (mm)</td>
<td>77.6</td>
<td>77.3</td>
<td>16.5</td>
</tr>
<tr>
<td>Losses (dB) at 1GHz</td>
<td>0.061</td>
<td>0.056</td>
<td>0</td>
</tr>
</tbody>
</table>
\[ A_{er} = \frac{P_r}{W_i} \]  \hspace{1cm} (5.11)

where \( P_r \) is the received power by the load and \( W_i \) is the power density of the incident plane wave. In terms of the de-embedded mixed mode power waves, the received power corresponds to the response in the differential port

\[ P_r = \frac{|b'_d|}{2} \]  \hspace{1cm} (5.12)

and the power density can be calculated as

\[ W_i = \frac{G_{AMP}G_T|a_d|^2}{4\pi d^2} \]  \hspace{1cm} (5.13)

where \( d \) is the distance between the antennas, and \( G_T \) and \( G_{AMP} \) are the transmitter antenna and amplifier gains, respectively. Replacing (5.12) and (5.13) in (5.11) and replacing the power waves by the de-embedded mixed transmission parameter \( S'_{ds} \) of the \( S_{Link} \) matrix, the realized effective area yields

\[ A_{er} = \frac{4\pi d^2}{G_{AMP}G_T} |S'_{ds}|^2 \]  \hspace{1cm} (5.14)

Now, the gain of an antenna can be calculated as [21]

\[ G = \frac{4\pi A_{er}}{\lambda^2 \tau} \]  \hspace{1cm} (5.15)

which, replacing the effective area by (5.14) yields

\[ G = \left( \frac{4\pi d}{\lambda} \right)^2 \frac{|S'_{ds}|^2}{\tau G_{AMP}G_T} \]  \hspace{1cm} (5.16)

where \( \tau = 4R_{in} \text{Re}(Z_{0diff}/|Z_{in} + Z_{0diff}|)^2 \) is the transmission coefficient, which depends on the impedance matching between the AUT and the VNA, and \( \lambda \) is the wavelength.

Therefore, the gain can be calculated directly from the mixed mode transmission parameter \( S'_{ds} \). Unlike the expression for the differential gain presented in [77], the formula developed in this work does not require estimating the components of the electric field produced by each arm of the differential antenna. The gain can be directly calculated knowing the mixed S parameters and the characteristics of the experimental setup.
5.6 Results

The proposed procedure was used to characterize three wire antennas and five samples of inert EEDs in the frequency range of 200 MHz to 3 GHz.

5.6.1 Differential Antennas

The input impedance and the gain of two dipole antennas and a V antenna were measured. The experimental setup inside the anechoic chamber is shown in Fig. 5.5. The antennas were built with AWG21 wires. The feed point consisted of two SMA male pins soldered to each arm of the wire antenna. Figs. Fig. 5.6-Fig. 5.8 show the measurements of two 10.5-cm and 21.3-cm long dipoles and a 20-cm V antenna, respectively. In the same figures, simulation results using the TDIE technique are presented. It can be seen that the measurements are in very good agreement with the numerical results, in the complete frequency range. The variation in the measured gain is due to the fact that the gain of the reference antenna was assumed constant and equal to its mean value for the complete frequency range.

![Characterization of the reception parameters of a dipole.](image)

![Graphs showing measured and simulated impedance.](image)
Fig. 5.6. a) Impedance and b) gain in the broadside plane of a 10.5 cm length dipole. Diameter=0.7 mm.

Fig. 5.7. a) Impedance and b) gain in the broadside plane of a 21.3 cm length dipole. Diameter=0.7 mm.
Fig. 5.8. a) Impedance and b) gain in the broadside plane of a V antenna. Length=20cm, diameter=0.7 mm, Angle between arms =90°.

5.6.2 Measurements of EED impedances

The differential input impedance of five inert EEDs was measured using the balanced technique. The characterization was performed using the differential adapter above described and two ports of the VNA, see Fig. 5.9.

The only difference between an inert EED and a (normal) EED is that the propagation medium in the last section of the transmission lines is air instead of lead azide. These measurements were used to verify the transmission line model detailed in Chapter 2, which is an analytically way to calculate the impedance of EEDs. Fig. 5.10 shows the dimensions of a characterized device and the electrical properties of the EED fillers used for the analytical model. Losses and permittivity of rubber were based on possible values of this material in the microwave range [78, 79]. As shown in Fig. 5.11, a good agreement is obtained between measurements and calculations. A difference of 3.42% in the value of the first resonant frequency is obtained.
To observe the variability of the EED impedance, two more empty inert EEDs and two filled inert EEDs were characterized. The devices were purchased from Securesearch, Inc. and both types have the same dimensions and external materials. In the empty EED, the only material inside the cap is rubber, which provides lead-in wires support. On the other hand, the filled one includes fillers in place of the explosives materials to be x-ray correct. These fillers are a small lead plug in the center, representing the lead azide, and nylon, representing the base charge.
Fig. 5.12. (a) Real and (b) imaginary parts of the measured input impedance of empty and filled inert EEDs. The impedances calculated from two analytical models are also plotted.

Fig. 5.12 presents the measured input impedance of two empty and two filled inert EEDs. All devices present a clear resonance in the measured frequency range (0.5 – 3 GHz). The figure shows that the input impedance response of the empty devices is very close. On the other hand, the response of the filled devices is more variable since a difference of 8.6% in the first resonant frequency was obtained. In addition, the first resonance frequency of the filled inert EED was smaller as compared to that of the empty. A reduction of 10% (19 MHz) was presented for the most distant case. It was noted that variations (in the order of milimeters) in the lead-in wires length could produce response differences as the observed.

Two analytical models, based on the real dimensions of the measured inert EEDs, were used to calculate their input impedance. The models, whose characteristics are summarized in Table 5.2,
intended to replicate the behavior of the empty and the filled devices. Variables in Table 5.2 are depicted in Fig. 5.13. Fig. 5.12 shows that a good agreement between the calculated and the measured inert EED input impedance is obtained. The resonance frequency predicted by the model is in the range of the measured one. Although the amplitude of the peak impedance real part was not well predicted by the models, it is shown that the measured amplitude is between the obtained with the empty and the filled models.

Table 5.2. Characteristics of the inert EED analytical models

<table>
<thead>
<tr>
<th>Model</th>
<th>$z_1$ (mm)</th>
<th>$z_2$ (mm)</th>
<th>$z_3$ (mm)</th>
<th>$D$ (mm)</th>
<th>$s$ (mm)</th>
<th>$\varepsilon_{r1}$</th>
<th>$\tan\delta_{1}$</th>
<th>$\varepsilon_{r2}$</th>
<th>$\tan\delta_{2}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Empty</td>
<td>8.0</td>
<td>14.5</td>
<td>2.0</td>
<td>7.0</td>
<td>3.0</td>
<td>2.0</td>
<td>0.01</td>
<td>1.0</td>
<td>0.001</td>
</tr>
<tr>
<td>Filled</td>
<td>8.0</td>
<td>14.5</td>
<td>2.0</td>
<td>7.0</td>
<td>3.0</td>
<td>2.0</td>
<td>0.1</td>
<td>1.0</td>
<td>0.0012</td>
</tr>
</tbody>
</table>

Fig. 5.13. Schematic of an EED including the variables used in Table 5.2.

### 5.7 Conclusions

The mixed mode S parameters were used to characterize experimentally the input impedance and gain of differential devices. The expressions to calculate these parameters directly from the mixed mode S parameters were derived. The de-embedding of the jig adapter, including its pins, was performed by modeling it with sections of transmission lines. It was shown that with this technique and with an adequate de-embedding of the adapter, an accurate and wideband characterization of differential antennas is possible.

This technique was also applied to measure EEDs input impedance. Stable and accurate measurements of five inert EEDs were achieved. These results were used to compare an analytical model of EED, obtaining a good agreement between measurements and calculations.

The measurement technique here described is applied in the next chapter to characterize the receiving properties of a sample of EED firing circuits. This characterization is used to calculate the mean received power of an EED connected to a firing circuit due to an external CW radiated field.
Chapter 6
Electromagnetic Characterization of Inert Improvised Explosive Devices*

Recently, EEDs have been used in the fabrication of inert improvised explosive devices (IEDs). The remote detonation of these devices has been proposed due to the susceptible response of EEDs to EM disturbances. The experimental characterization of the electromagnetic response of a set of IEDs is presented in this chapter. The characterization is based on measurements of gain and input impedance of the firing circuits of inert IEDs in the range of 200 MHz to 2 GHz. First, a theoretical model to predict the average gain is developed. A comparison between the measurement results and theoretical predictions is presented. Then, the power delivered to a hot wire based EED connected to the inert IED is calculated for each measured case and it is statistically analyzed. The effects on the IED response of different characteristics of the coupling, such as the IED’s elements, the incident field polarization, and the impedance mismatch, are also studied. In addition, the statistical performance of the induced power and current are determined and compared with the results of a numerical model based on Monte Carlo simulations. It is demonstrated that both approaches, a theoretical model based on an equivalent antenna and Monte Carlo simulations, can be used to describe the average electromagnetic response of the firing circuit of IEDs.

6.1 Introduction

Improvised explosive devices (IEDs) are unconventional bombs used in many regions in conflict of the world. These devices are handmade and have different shapes and methods of activation [3]. Their electric version is commonly composed of a dielectric container, shrapnel, explosive material, and an electric activation (firing) circuit. Its remote detonation using radiated energy has been a subject of interest in the last years. Although studies on the effects of intentional electromagnetic interferences (IEMI) on IEDs are recent [36, 63], works on the radiated susceptibility of this kind of circuits have been carried out earlier since the detonator used in these devices is commonly employed in the industry [8, 14, 16]. The activation mechanism of electrical IEDs is based on

feeding a hot wire based electro-explosive device (EED) with an electrical current to increase the
temperature of its bridge-wire until a critical value. The feeding current is usually produced when a
switch is closed and, as a consequence, the EED is connected to a battery. Remote activation could
be achieved if an IEMI induces a current with sufficient energy in the EED feed wires.

To describe the suitable electromagnetic (EM) properties for a remote activation, a
characterization of the IED’s EM behavior is necessary. A first study was carried out by Lambrecht
et al., who developed an analytical model for the input impedance of an EED [80]. Based on this
approach, some strategies to model the EM coupling to EEDs with wires were proposed, as
explained in Section 2.2 and in [25, 81]. Due to their handmade fabrication, IEDs have variable
designs and random physical properties. The effect of the variability of the wiring on the EM
response was discussed in Chapter 4 (see also [63, 82]). These studies have shown that remote
detonations can be achieved with existing radiation sources and have allowed the determination of
the optimal frequency for a continuous wave excitation.

In this chapter, an experimental EM characterization of the firing circuit of inert IEDs is
presented and compared with theoretical and numerical models. The physical characteristics,
excluding the explosive fillers, and the inherent variability of the IEDs were replicated in the 22
characterized samples.

6.2 Test Description

6.2.1 Samples

A group of 22 IEDs† were experimentally characterized. To facilitate the manipulation and the
identification of the influence of the firing circuit variability on the EM response of IEDs, inert
devices were used. The inert IEDs were replicas of actual IEDs, but without detonator or explosive
load. Each sample was made by different explosives technicians with the knowledge of the
manufacturing techniques of IEDs used in Colombia by illegal armed groups. Each technician had
freedom in choosing the IED type, materials, and dimensions. As a result, the samples featured the
typical IED random parameters.

Since the study is focused on devices susceptible to electromagnetic radiation, only electrically
activated IEDs were considered. The samples were classified in three groups according to their
switch structure (see Table 6.1).

---

† IEDs were provided by Mr. Ernesto Neira from the Escuela de Ingenieros Militares de Colombia.
Fig. 6.1. Histograms of the connection wire lengths of the measured samples and their fits to a Gamma distribution. Parameters $a$ and $b$ are, respectively, the shape and scale parameters of the Gamma distribution.

Fig. 6.2. Schematic of the firing circuit of an electrically activated IED.

All the samples were built with the same electrical components connected in series that included: a switch, a battery and connection wires. Except for the detonator, this configuration represents the typical circuit used in real electrical IEDs. In addition, two syringe type samples featured special characteristics: i) one sample included shrapnel and ii) another included twisted pair wires. The samples presented containers and connection wires of different dimensions.

Table 6.1. Classification of the Characterized IEDs

<table>
<thead>
<tr>
<th>Type</th>
<th>Syringe</th>
<th>“M”</th>
<th>Sponge</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Photograph</strong></td>
<td><img src="image1" alt="Syringe" /></td>
<td><img src="image2" alt="M" /></td>
<td><img src="image3" alt="Sponge" /></td>
</tr>
<tr>
<td><strong>Description</strong></td>
<td>The switch of the circuit is a syringe activated by pressure. The electric contacts are made with the wires.</td>
<td>The switch of the circuit is made of an M-shape metallic sheet. When the sheet is bent, the IED is activated.</td>
<td>The switch of the circuit is made of a sponge activated by pressure. The electric components are in the interior of the container.</td>
</tr>
<tr>
<td><strong>Container Material</strong></td>
<td>PVC, Plastic, glass</td>
<td>PVC, Plastic, Tinplate</td>
<td>Tinplate</td>
</tr>
<tr>
<td><strong>Quantity</strong></td>
<td>14</td>
<td>7</td>
<td>1</td>
</tr>
<tr>
<td><strong>Trigger mechanism</strong></td>
<td>Pressure, victim-activated</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td><strong>Use</strong></td>
<td></td>
<td></td>
<td>Antipersonnel mine</td>
</tr>
</tbody>
</table>
Cylindrical containers with 13.1 cm of average height and 97.8 cm$^2$ of average base area were used. The histogram of the length of each arm of the firing circuit wiring is shown in Fig. 6.1. Arm A corresponds to the direct connection between an EED pin and the switch, while arm B is the connection that includes the battery, as shown in Fig. 6.2. The total length of the connection wires was 66.8 cm in average.

The main interest in this study is to describe the far field EM response of IEDs. Since devices without explosive loads filling the container were used, the effect of the inclusion of these materials on the far field receiving properties was investigated. From full-wave simulations using typical explosives, it was verified that this effect on the gain can be neglected. On the other hand, it was found that the explosives permittivity modifies the wiring effective length. However, it will be shown that due to the randomness of the devices, the effect of the presence of explosives on the mean input impedance and mean induced power is not significant. In addition, the simplification allows focusing the analysis on the firing circuit variability.

### 6.2.2 Experimental Setup

In order to measure the IEDs capability to receive radiated electromagnetic energy, two typical properties of antennas were measured: gain and input impedance. To consider the IED’s firing circuit as a receiving antenna, let us determine the factors that mainly affect its susceptibility to an incident electromagnetic field and identify the range of frequencies at which the best coupling can be achieved.

The gain and the input impedance of inert IED samples were measured by placing the instrument port at the position where the EED would be connected. As a consequence, the input port of each inert IEDs is a two-wire transmission line. The characterization of devices with non-standard high-frequency connectors, such as two-wire transmission lines, is usually made using baluns, which have a limited frequency response and can affect the measurement accuracy. For this reason, the technique to fully characterize (to measure the input impedance and gain) differential devices presented in Chapter 5 was used.

The measurements were performed using a 4-port vector network analyzer (VNA) inside the anechoic chamber of the Universidad de los Andes. The input impedance $Z_{\text{in}}$ and the gain $G$ of the inert IEDs were obtained using (5.10) and (5.16), respectively. As reference antenna, a Log-periodic antenna model 3148B from ETS-Lindgren was used. The measurements were performed in the range of 200 MHz to 2 GHz.
6.2.3 Procedure

Each sample was illuminated from four different angles of incidence and with the reference antenna in both polarizations. As a result, eight measurements were performed for each sample. Fig. 6.3 shows the four angles of incidence. In summary, the variations of the angle of incidence and the antenna polarization were:

- Azimuth: $\phi = 0^\circ, 90^\circ$ and $270^\circ$.
- Elevation: $\theta = 0^\circ$ and $90^\circ$.
- Antenna polarization: horizontal and vertical.

Since 22 samples were characterized, 176 measurements were performed in total.

6.3 Theoretical Model

6.3.1 IED in Reception

Different approaches have been employed to study the electromagnetic susceptibility of devices that include EEDs in their circuits. A common model, used for different applications, is based in the representation of the firing circuit as an antenna and the EED as its (lumped) load [8, 14, 16, 63, 82]. Using this concept, Thévenin and Norton equivalents or antenna theory can be used to calculate the received power in the EED due to an incident plane wave. Applying the second approach, the received power in the frequency domain yields [21]

$$P_d = W_i \tau \frac{\lambda^2 G}{4\pi}$$  \hspace{1cm} (6.1)

where $W_i = |E_i|^2/(240\pi)$ is the incident power density, $E_i$ is the incident electric field, $\tau = (1 - |\Gamma_L|^2)$ is the transmission coefficient (also known as the reflection mismatch efficiency [83]), $\Gamma_L = (Z_{EED} - Z_A^*)/(Z_{EED} + Z_A)$ is the reflection coefficient between the antenna (IED wiring) with impedance $Z_A$ and the load (EED) with impedance $Z_{EED}$, and $G$ is the antenna gain.
All the variables in (6.1) are frequency dependent. From this expression, it is possible to see the parameters that determine the frequency response of the received power and, as a consequence, the behavior of the coupling between an incident wave and an IED. Considering a constant excitation, the received power becomes a function of the receiver parameters (τ and G). Thus, maximizing (6.1) will result in the frequencies at which the system is more susceptible.

In a deterministic problem, τ, G, and the frequency of maximum coupling (f_{opt}) have deterministic values. However, if the coupling has random properties (e.g. random lengths and geometries of the firing circuit, random device’s dimensions, or unknown angle of incidence), the process becomes stochastic. In these cases, only a statistical description of f_{opt} is possible.

### 6.3.2 Mean Effective Gain

Mean effective gain (MEG) is a figure of merit commonly used to compare the performance of antennas in environments with random properties, such as multipath environments without line-of-sight (LOS) [84]. In that case, there are multiple incident signals with unknown angle of incidence.

#### 6.3.2.1 Single Incidence with Line-of-Sight

The coupling of a single incident wave to a receiver can also be characterized by a simplified version of the MEG. This is a scenario with LOS and with no random field components; then, the MEG, G_e, becomes equal to the receiver gain in the direction of the LOS expressed in terms of its components [85]:

\[
G_e = \frac{1}{1 + \text{XPR}} \left( \sqrt{\text{XPR}
G_{\theta}(\theta_0, \phi_0) + \sqrt{G_{\phi}(\theta_0, \phi_0)}} \right)^2, \tag{6.2}
\]

where XPR = P_V/P_H is the cross-polarization power ratio of the incident wave, P_V and P_H are the mean incident powers of the vertically and horizontally polarized incident waves, respectively, and G_\theta and G_\phi are the \theta and \phi polarized components of the antenna power gain, respectively.

XPR depends on the excitation polarization; as a result, it can be considered as a deterministic quantity. In an IED in reception, the gain components are random variables since the properties of the IEDs are not known.

#### 6.3.2.2 Single Incidence with Line-of-Sight and a Random Receiver

Now, consider the MEG variability due to the change of the receiver. It will generate different MEGs for different receivers, which can be statistically characterized by their average value. Here, it is proposed that this average behavior can be replicated by an equivalent structure. Therefore, the
MEG of an IED, for a single incident wave with equal polarization components (XPR = 1), can be calculated using (6.2) as

\[ G_e(\omega) = \frac{G_d(\omega, \theta, \phi)}{2} \]  

(6.3)

where \( G_d(\omega) \) is the gain of the equivalent structure as a function of the frequency.

6.3.2.3 Equivalent Structure of an inert IED

In [86], it is shown that the resonant properties of a small wire antenna with arbitrary geometry are mainly determined by the effective length and the effective volume, and not by the total wire length and geometry. In addition, the radiation patterns are only established by one component (vertical) of the current; then, the pattern of different wire antennas over a perfect ground plane converges to that of a straight-wire monopole. Although this analysis is made for small antennas, the same conclusions are extended to electrically longer antennas beyond the second resonance [87]. Thus, the behavior of an arbitrary antenna can be replicated with other antennas of similar effective length and volume [87].

Using this concept, it is possible to represent the average performance of a set of IEDs’ firing circuits by means of a simplified structure that maintains the average effective dimensions. To replicate the behavior of the IEDs, that structure is restricted to: i) be a wire antenna, ii) cover a cylindrical volume with the mean effective dimensions of the samples, iii) have a low gain, and iv) have a constant and real input impedance between 0.5 GHz and 2 GHz. The last condition was established based on the measured samples. It was observed that as the number of samples increases, the average values of the imaginary and real parts of the impedance of these devices converge to zero and to 280 \( \Omega \), respectively, for frequencies above 0.5 GHz.

An antenna that fulfills the required characteristics is the double helix. Starting from the average physical dimensions of the measured samples, the double helix was modified until achieving a constant real input impedance in most of the required frequency range. The obtained structure, shown in Fig. 6.4, has a base radius of 10 cm, a top radius of 5cm, a height of 20 cm, and 2 turns. The resulting total wire length was 202.7 cm.

In addition, the container’s influence was modeled using a plane-wave analysis for dielectric containers and a shielding calculation for the metallic ones. The plane wave analysis, commonly used to model antenna radomes, assumes that the dielectric barrier is locally a planar slab. Thus, the slab transmission coefficient \( T \), which is the ratio between the transmitted and incident electric fields, can be calculated for a normal incidence as [88] (Ch. 53, pp. 5-6)
Fig. 6.4. Geometry of the Double Helix antenna.

\[ T = \frac{2}{A + B + C + D} \]  \hspace{1cm} (6.4)

where the unknowns are the elements of the slab ABCD matrix

\[
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix} = \begin{bmatrix}
\cosh(jk_x t) & Z_r \sinh(jk_x t) \\
Z_r^{-1} \sinh(jk_x t) & \cosh(jk_x t)
\end{bmatrix}
\] \hspace{1cm} (6.5)

\( t \) is the slab thickness, \( k_x \) is the propagation constant, \( Z_r = \sqrt{\mu_r/\varepsilon_r} \), and \( \mu_r \) and \( \varepsilon_r \) are the slab’s complex relative permeability and permittivity, respectively.

For metallic containers, it was assumed that only the incident waves impinging by the switch aperture (vector \( k_3 \) in Fig. 6.3) couple with the IED wiring. For this reason, the average transmission coefficient of these containers was approximated to \( 1/4 \) in the complete frequency range since only one of the four incidences contributes to the mean gain. Table 6.2 presents the characteristics of the containers and \( T \) calculated for each material at 2 GHz. Typical values of the permittivity and tangent losses in the microwave range were used.

The average attenuation due to the containers was calculated as the pondered average of the \( T \) values of each material as a function of the frequency. Thus, \( G_d \) in (6.3) can be replaced by the gain of the equivalent double helix antenna multiplied by container’s average attenuation.

<table>
<thead>
<tr>
<th>Table 6.2: Characteristics of the IED Containers</th>
</tr>
</thead>
<tbody>
<tr>
<td>Plastic</td>
</tr>
<tr>
<td>Samples Quantity</td>
</tr>
<tr>
<td>( \varepsilon_r )</td>
</tr>
<tr>
<td>( \tan\delta )</td>
</tr>
<tr>
<td>Thickness (mm)</td>
</tr>
<tr>
<td>( T ) (dB) at 2 GHz</td>
</tr>
</tbody>
</table>
6.3.3 Average MEG: Multiple Incidences with Line-of-Sight and a Random Receiver

To include multiple angles of incidence in the theoretical MEG, an average MEG can be calculated as

\[
G_{ave}(\omega) = \frac{1}{N} \sum_{i=1}^{N} G_{ei}(\omega) \tag{6.6}
\]

where \(N\) is the number of measured/computed incidence angles and \(G_{ei}\) is the MEG for each incidence. It is important to note that when the number of considered incident angles increases (i.e. \(N \rightarrow \infty\)), the sum in (6.6) converges to the typical integral form of the MEG presented in [84] and the angular power density functions of the incoming waves should be used. Using that integral, it can be shown that, if the incident angles have a uniform distribution, the average MEG is independent of the gain and becomes equal to half its radiation efficiency [84].

6.4 Numerical Simulation

Monte Carlo method is a numerical technique to model the response of complex systems with random parameters. This method is used in this study to obtain the EM response of the IED’s firing circuit, which has variable wiring geometry. Since any radiated disturbance will mainly couple to the wiring of the IED, a simplified representation composed by the wiring and the critical load (i.e. EED) was considered.

To calculate the induced power in the EED, the wires and the EED were independently modeled. First, the reception properties of the wires were calculated. The gain and the wire’s input impedance at the connection point of the load were obtained using the Time Domain Integral Equation (TDIE) technique implemented with a marching-on-time scheme [64]. Then, the impedance of the EED was

![Diagram of angles for the generation of arbitrary wire geometries](image)

Fig. 6.5. Angles for the generation of arbitrary wire geometries. Inclination angles \(\phi_{1A}, \phi_{2A}, \phi_{1B}, \phi_{2B}\) are measured in the XY-plane (a) and inclination angles \(\theta_A\) and \(\theta_B\) are measured in the YZ-plane (b).
calculated using an analytical model [80]. Finally, these parameters were inserted into (6.1) to obtain the induced power as a function of the incident electric field.

This way of determining the device response from its elements’ responses independently calculated, allows reducing the computation time, since only the variability of one element is considered. In this case, the wire geometry was varied randomly while the load and the incident field were kept fixed.

The geometry of the simulated structures was based on the physical characteristics of the IEDs described in Section 6.2.1. The wires were allowed to change in three dimensions by modifying its length and pattern. Arbitrary patterns were obtained by dividing the wire in two arms, each one with two straight sections as shown in Fig. 6.6. The critical load was placed in the union of both arms. The length of the sections of each arm is half of the arm length, which is a random variable with a gamma distribution. The wire pattern was modified assigning random uniformly distributed inclination angles at each section as illustrated in Fig. 6.5. The probability distribution parameters of each variable are presented in Table 6.3.

400 arbitrary wire structures were simulated using TDIE. A wire’s radius of 1.4 mm was used. The incident electric field was simulated with magnitude of 1 V/m and direction of propagation \( \hat{x} \) and polarization \( \hat{y} \), as shown in Fig. 6.6. For each case, the gain for \( \phi = 270^\circ \) and \( \theta = 90^\circ \), and the wire’s input impedance were calculated.
Table 6.3. Parameters of the EM Coupling Simulation

<table>
<thead>
<tr>
<th>Variable</th>
<th>Description</th>
<th>Distribution Parameters*</th>
</tr>
</thead>
<tbody>
<tr>
<td>L_A</td>
<td>Length of Arm A</td>
<td>Gamma 3.17 7.13 cm</td>
</tr>
<tr>
<td>L_B</td>
<td>Length of Arm B</td>
<td>Gamma 5.81 7.62 cm</td>
</tr>
<tr>
<td>φ₁A, φ₁B</td>
<td>Inclination angles φ₁A and φ₁B</td>
<td>Uniform 20° 160°</td>
</tr>
<tr>
<td>φ₂A, φ₂B</td>
<td>Inclination angles φ₂A and φ₂B</td>
<td>Uniform -90° 90°</td>
</tr>
<tr>
<td>θₐ, θᵦ</td>
<td>Inclination angles θₐ and θᵦ</td>
<td>Uniform 20° 160°</td>
</tr>
</tbody>
</table>

*For the Gamma distribution, \( a \) and \( b \) are the shape and scale parameters, respectively. For the Uniform distribution, \( a \) and \( b \) are the minimum and maximum values, respectively.

6.5 Electromagnetic Response of IEDs

In this section, the measured data are presented and analyzed. First, the gain and the input impedance of the firing circuit of the IEDs are described. Then, the dissipated power in an EED connected to the inert IEDs is calculated analytically using the measured parameters. Using this information, the main factors that describe the coupling with the system are determined.

6.5.1 Gain

Fig. 6.7 shows the broadside (incidence k4 in Fig. 6.3) gain measured for two samples. For these cases, the reference antenna is in vertical polarization (\( \hat{z} \) direction in Fig. 6.3). The gain measured for a syringe-type sample, shown in Fig. 6.7a, presents a behavior close to the one of a dipole; the gain strongly decreases in specific (anti-resonant) frequencies. Other samples show a more random behavior, presenting some narrowband peaks in the gain. This is probably due to influence of the IED circuit and casing elements in the near field and the current distribution. Fig. 6.7b, for example, shows the gain of a sample with a metallic container and presents a pronounced frequency selective behavior.

The gain of a syringe-type sample with a plastic container for different incidences is shown in Fig. 6.8. As shown in the figure, there is an angle of incidence and polarization that provides a positive gain (in decibels) in most of the frequencies. It corresponds to a broadside vertically polarized incident wave when the device is vertical, similar to a vertical dipole (curve with circular marker in Fig. 6.8). Changing the polarization to horizontal and changing the incidence to a top incidence (vector k3) produces a considerable reduction in the gain in frequencies bellow 1 GHz.
Fig. 6.7. Broadside gain of a syringe-type sample with a glass container (a) and a sponge-type sample with a metallic container (b).

Fig. 6.8. Gain of a syringe-type sample with a plastic container for different angles of incidence and antenna polarizations.

Fig. 6.9. Mean measured gain compared with predictions using (5).
For higher frequencies, higher modes in the structure are formed and positive gains are obtained for all the cases. This interesting behavior is representative of most of samples with plastic container.

To describe the gain behavior of all the samples, the average of the measured gains was calculated for each frequency. Fig. 6.9 shows the mean measured gain. It presents a general increasing dependence with the frequency. It is compared with the prediction of the double helix model using (8). The figure shows that both gains vary in the same range of values.

6.5.2 Input Impedance

The input impedance of the firing circuit of an IED shows the self-resonant frequencies and determines the frequencies at which it could be matched to a specific load. Fig. 6.10 shows the input impedance and the resonant behavior for the syringe-type and sponge-type samples whose gains are presented in Fig. 6.7. The resonance frequencies can be identified as the frequencies at which the imaginary part of the impedance is zero. The resonances depend on the physical characteristics of the samples; in fact, the wire’s effective length is one of most dominant factors.

This resonant behavior produces a strong variability of the impedance as a function of the frequency.
frequency and establishes a number of frequencies of maximum power transfer to a determined load. In the case of the inert IED samples, the load corresponds to an EED that also presents a resonant behavior in this range of frequencies. Fig. 6.11 shows the typical input impedance of an EED calculated using the analytical model SABC (State of the Art Blasting Cap) [80]. The loss of power as a function of the frequency due to this impedance mismatch is quantified by the transmission coefficient $\tau$.

### 6.5.3 Mismatch Effect

From (6.1), it is evident that the impedance mismatch between the receiver and the load affects the value of the received power. Since in an IED the impedances of both elements are frequency dependent, the resulting transmission coefficient $\tau$ is also frequency dependent and present maximum values at the frequencies at which the impedances are complex conjugates of each other, obeying the maximum power transfer theorem [89].

The box plot and the mean value of the transmission coefficient $\tau$ as a function of the frequency are presented in Fig. 6.12. This parameter was calculated using the measured impedance of the inert IEDs and the typical input impedance of an EED shown in Fig. 6.11. Fig. 6.12 shows that, in general, there is a considerable mismatch in the measured frequency range. However, there is a range, at about 0.761 GHz, at which the mean value of $\tau$ increases and reaches its maximum value of 27%. This frequency corresponds to the best average impedance matching. This figure also shows that for some cases, represented by the minimum values of $\tau$, the losses due to mismatch are severe. For instance, the transmission coefficient at the best coupling frequency can have a

![Box plot showing transmission coefficient vs frequency](image)

Fig. 6.12. Box plot of the transmission coefficient associated with the mismatch between the characterized inert IEDs and a typical EED. On each box, the central mark is the median, the edges of the box are the lower and upper quartiles, the whiskers extend to the most extreme data points, and the circles are the outliers. The mean value and the transmission coefficient obtained with the double helix model are also plotted.
minimum value as low as 1.1%.

As with the gain, the mean response of $\tau$ is compared with the response of the double helix model. Fig. 6.12 shows that a good agreement between the model and the mean response is obtained. In particular, both (mean and double helix) predict the same frequency at which the transmission coefficient reaches its maximum values. Notice that these maxima occur close to the EED anti-resonant frequencies. In fact, in these frequencies, there is a higher probability of achieving the maximum power transfer condition due to the high variability of the EED impedance and to the flat frequency response of the mean impedance of the samples.

### 6.5.4 Power Dissipated in an EED

It is worth reminding that the characterized samples are prototypes of inert IEDs. The measurement port was the point at which the EED is normally located. Consequently, if the measured inert IED’s gain and input impedance and an EED’s impedance are inserted in (6.1), it is possible to calculate the power delivered to the EED. Thus, the activation of the IED prototype due to an incident plane wave can be determined.

Fig. 6.13 presents the power received by a typical EED used as the load of the characterized inert IEDs. In the measured frequency range, a variation of the mean power higher than 10 dB is obtained. The figure shows that there is a frequency at which the mean bridge-wire power is maximum. This frequency corresponds to $f_{opt} = 0.761$ GHz and represents the frequency of the

![Box plot of the received power in an EED as a load of the measured inert IEDs due to an incident plane wave of $|E_0| = 1$ V/m. The mean value and the power using the double helix model are also plotted. See the box plot explanation in Fig. 6.12.](image-url)
average best coupling. In this case (i.e. with the impedance of the EED presented in Fig. 6.11), this frequency coincides with the value obtained in the previous section for the transmission coefficient \(\tau\). From Fig. 6.9, Fig. 6.12, and Fig. 6.13, it is shown that \(\tau\) is the dominant factor in the frequency response of the IEDs (for the considered EED impedance). In general, the response of the mean induced power depends on the combined contribution of the gain and the transmission coefficient \(\tau\).

Observing the box plot for each frequency point, Fig. 6.13 also illustrates the significant variation (30 dB or higher) of the received power due to the variability of the samples. Interestingly, the dispersion is highly unsymmetrical: most of the power values are observed to be near the upper limit, as can be seen from the curve presenting the mean value and the box plot.

To compare the induced power inferred from the measurements and the one obtained from the theoretical model, the power dissipated in an EED attached to the double helix model is also plotted in Fig. 6.13. This power was calculated replacing the double helix’s MEG and \(\tau\) in (3). As expected, the general behavior of the mean received power is followed. The power values as well as the frequency dependence are well predicted.

The best coupling frequency represents the frequency at which the characterized IEDs are more susceptible in average. A disturbing incident field tuned at this frequency will induce a higher power on average compared with a signal tuned in another frequency in the measured range. As shown earlier, the theoretical model can be used to predict the frequency behavior of the received power in the IED and the best coupling frequency. However, it should be noted that the model is capable of describing the mean behavior but not other statistical moments. A more detailed statistical description of the EM coupling was developed with Monte Carlo simulations and is presented in the next section.

### 6.6 Statistical Properties of the Coupling

In this section, the results of the Monte Carlo simulations (MCS) are compared with the measurements. First, the induced power in the critical load is presented as a function of the frequency. Then, the probability functions of the involved variables are deduced. Finally, the probability of activating an IED for an incident electric field with a given intensity and at a specific frequency is evaluated by means of the survivor function.

#### 6.6.1 Dispersion of the Induced Power

The power delivered to an EED as a load of the simulated wirings was calculated following the procedure described in the previous section. Fig. 6.14 presents the mean power received in the
critical load calculated using the measurements and using the MCS. Both results show a similar dependence of the power with the frequency. Despite the simplicity of the model used in the MCS, it is shown that it predicts well the power levels and the frequency ranges at which the mean power increases. A maximum value close to 0.7 GHz is presented in both curves.

To illustrate the data dispersion, Fig. 6.15 shows the standard deviation of the received power as a function of the frequency. Both results, based on MCS and based on measurements, show an overall decrease with the frequency. It is observed that this dependence is weaker beyond 1 GHz, especially for the MCS case. An average value of the standard deviation around $\sigma = 8$ dBm is obtained for both approaches in the considered range of frequencies.

Using the standard deviation, it is possible to calculate the probability of having the mean value of the power inside a confidence interval (CI) for each method. That is

$$P_{CI} = P_r(\bar{P}_d - \mu \leq \text{CI}).$$

If we define the standardized variable $Z = \frac{n^{1/2}}{\sigma} (\bar{P}_d - \mu)$; then, the probability can be expressed as a function of the standard deviation and the number of samples, $n$. [90]
In the next section, it will be shown that the induced power in logarithmic scale follows an extreme value (EV) distribution. Using the standard EV distribution and the number of samples in (11) yields the probability of the CI for each method. Table 6.4 shows that the probability for a CI of 1.5 dB is higher than 92% in both cases.

<table>
<thead>
<tr>
<th>Method</th>
<th>CI (dB)</th>
<th>n</th>
<th>P_CI (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Monte Carlo</td>
<td>1.5</td>
<td>400</td>
<td>97.7</td>
</tr>
<tr>
<td>Measurements</td>
<td>1.5</td>
<td>176</td>
<td>92.0</td>
</tr>
</tbody>
</table>

### 6.6.2 Probability Distribution Functions

To analyze in detail the dispersion of the induced power, the probability distribution function (PDF) was calculated. Using the statistics toolbox of MALTAB, it was determined that the best fit of the PDF for the power in dBm is obtained with an extreme value (EV) distribution (also known as Gumbel distribution). This result agrees with the distribution of the induced power in a uniform multipath environment without line of sight [88] (Ch. 58, pp. 3), which is a canonical scenario of coupling with random properties. Fig. 6.16 presents the histogram of the data and the fitted EV distribution for the power induced in the EED at 0.761 GHz.

Fig. 6.17 presents the histograms of the power in watts and the current magnitude and their respective distributions. These were fitted in MALTAB and it was verified that the transformation of distribution agrees with the analytical random variable transformation. In addition, using the Kolmogorov-Smirnov test with a 5% of significance level, it was verified that these distributions describe well the behavior of the variables in all the range of frequencies.

### 6.6.3 Survivor Function

Assuming a linear response of the IEDs, it is possible to use the obtained received power to calculate the intensity of an incident field able to activate each sample. This can be calculated considering the PDF of the induced power due to the incident field and the cumulative distribution function (CDF) of the EED’s firing power threshold. This CDF was calculated from the distribution of the firing energy of a squib Mk 1 Mod 0 for adiabatic excitations presented in [91]. The equivalence between the firing energy and the firing power distributions was obtained by considering a constant power during a period of $\Delta t = 2.5$ ms. Since the considered period is small compared to the thermal constant of low energy EEDs, which ranges from 20 ms to 60 ms [81], the
heating process is adiabatic. The firing power distribution in dBm was obtained performing a variable transformation using the expression $P = 10 \log(W/\Delta t)$, where $W$ is the energy level. The energy firing distribution can be modeled using a Log-normal distribution [91]; as a consequence, the PDF of the firing power threshold expressed in dBm results in a normal distribution. Fig. 6.18 shows the empirical and fitted CFDs.

The firing probability $(FP)$, which represents the probability of having an induced power $P_i$ higher than the firing power threshold $P_T$, can be determined by means of a reliability analysis as [92, 93]

$$FP(E_{inc}) = p(P_i(E_{inc}) > P_T) = \int_0^{\infty} cdf_{P_T}(P)pdf_{P_i}(P) dP,$$  \hspace{1cm} (6.7)
where $cdf_{p_T}$ is the cumulative probability function of the firing power threshold and $pdf_{p_i}$ is the probability distribution function of the induced power. Replacing these functions, respectively, and expressing $pdf_{p_i}$ as a function of the incident field magnitude, $|E_{inc}|$, the firing probability yields

$$FP(E_{inc}) = \int_0^\infty \frac{1}{2} \left[ 1 + \text{erf} \left( \frac{P - \mu_T}{\sqrt{2}\sigma_T} \right) \right] \left[ \frac{1}{\sigma_i} e^{-\frac{(P - \mu_i)}{\sigma_i}} \right] e^{-e^{-\left( \frac{P - \log|E_{inc}| - \mu_i}{\sigma_i} \right)}} \] dP, \quad (6.8)$$

where erf is the error function, $\mu_T$ and $\sigma_T$ are the mean and the standard deviation of the $P_T$ normal fit, and $\mu_i$ and $\sigma_i$ are the location and scale parameters of the $P_i$ EV fit. $FP$ can be solved numerically and the survivor function (SF) can be obtained as $SF(E_{inc}) = 1 - FP(E_{inc})$.

Fig. 6.19 shows the survivor function obtained using the $pdf_{p_i}$ calculated with the parameters of the MCS results and the parameters from the measurements at 0.76 GHz. The figure shows that both methods reproduce the same behavior, but the MCS underestimates SF. In addition, it
illustrates that the quantity of activated devices for a given field intensity becomes a random variable due to the variability in the geometry and $P_T$. Thus, for example, an electric field of 12 kV/m and 2.5 ms is necessary to achieve a survival probability of 5%.

6.6.4 Power Levels of a Radiating Source

The information provided by the survivor function is useful to determine the power levels in a radiation source required for a remote activation. Consider an antenna with gain $G_T$ radiating a power $P_{TR}$ at a distance $d$ from the device. Assuming far field, the power necessary to produce a determined electric field $E_{inc}$ is given by [94]

$$P_{TR} = \frac{(E_{inc} d)^2}{60 G_T}. \quad (6.9)$$

Table 6.5 shows the necessary transmitted power to produce a remote activation as a function of the distance. An antenna with a gain of 20 dB and two electric field thresholds, which correspond to a survival probability of 10% and 5%, were used for calculations.

<table>
<thead>
<tr>
<th>$d$ (m)</th>
<th>$P_{TR}$ (MW)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$E_{inc} = 6$ kV/m $SF(E_{inc})=10%$</td>
</tr>
<tr>
<td>10</td>
<td>0.6</td>
</tr>
<tr>
<td>30</td>
<td>5.4</td>
</tr>
<tr>
<td>50</td>
<td>15.0</td>
</tr>
<tr>
<td>100</td>
<td>60.0</td>
</tr>
</tbody>
</table>

6.7 Conclusion

The electromagnetic response of a set of 22 inert IEDs with different designs, materials, and dimensions was characterized. Their capabilities of receiving EM waves, expressed in terms of their gain and input impedance, were measured, analyzed, and compared with a theoretical model and with Monte Carlo simulations. As a result, the resonant behavior of the IEDs was characterized.

The MEG concept was applied to this problem with deterministic incidence of excitation but random properties in the receptor. It was shown that the EM behavior of the average measured gain of a receiver with complex and random characteristics can be predicted by the average MEG of an equivalent receiver. In this case, a double helix antenna was used. In addition, it was shown that this equivalent structure could be used to describe the average behavior of the transmission coefficient.

The mismatch effect was evaluated using the transmission coefficient. It was shown that this parameter and the gain of the firing circuit determine the behavior of the coupling of an incident
EM plane wave with an inert IED. The dissipated power in an EED placed as a load of the inert IEDs was numerically calculated for each measurement. The mean dissipated power was obtained and the frequency of maximum average power dissipation was determined. It was shown that the double helix model and the Monte Carlo simulations predict well the frequency dependence of the dissipated power. Since more details are considered in the Monte Carlo simulations, a better description is obtained with this method.

The probability distribution functions of the induced power and current were deduced from the measurement results and from the Monte Carlo simulations. A good agreement between both approaches was observed. It was found that the induced power in dBm follows an Extreme Value distribution and that the power in watts and current magnitude follow Weibull distributions. Furthermore, the survivor function of the loaded samples as a function of the incident field magnitude was calculated for the best coupling frequency. It was determined that a continuous wave excitation of 12 kV/m is necessary to achieve a survivor probability of 5%.

In this study, inert devices without explosive material were used. In addition, they were placed in a free-space environment, even though in practice, they are partially buried. It is expected that the intensity levels of induced power will be modified by the properties of the IED’s fillers and the soil. Further research should include the effect of the permittivity and losses of these materials.
Chapter 7
Thermal Response of Electro-Explosive Devices

In previous chapters, transient and steady-state models of the thermal response of EEDs were used to relate induced power to bridge-wire temperature. In this chapter, these models are experimentally verified using a fiber optic thermometer coupled to the EED bridge-wire. Then, this assembly is used to measure the induced power in an EED due to an external electromagnetic field. A radiation test of two dipole antennas attached to inert detonators and three samples of inert improvised explosive devices (IED) with inert detonators is presented. The radiated energy was produced by a low power microwave (LPM) generator with output power of 800 W and operating frequency of 2.45 GHz.

7.1 EED Thermal Models

7.1.1 Transient Thermal Model
In Chapter 3, it was shown that the thermal response of an EED can be modeled by a first-order differential equation as

\[ p(t) - \gamma_{th} \theta_{BW} = C_{th} \frac{d\theta_{BW}}{dt}, \]  

(7.1)

where \( p(t) \) is the power dissipated in the EED’s bridge-wire as a function of the time, \( \gamma_{th} \) is the thermal conductance, \( C_{th} \) is the thermal capacitance, and \( \theta_{BW} \) is the bridge-wire temperature increase. This model is especially useful to describe the bridge-wire temperature due to transient excitations.

7.1.2 Steady-State Model
In steady-state, a simplification of (7.1) can be done ignoring the time derivative component in the equation. Thus, the temperature increase yields

\[ \theta_{BW} = P / \gamma_{th}, \]

or

\[ \theta_{BW} = R_{th} P, \]  

(7.2)
where $R_{th}$ is the EED’s thermal resistance, as explained in Chapter 4. In the next section, the linear relationship in (7.2) is experimentally verified by measuring the bridge-wire temperature increase for different power levels. From these curves, $R_{th}$ of different EEDs is characterized.

### 7.2 Temperature Measurements

#### 7.2.1 Dynamic Response

The thermal response of four EEDs was characterized measuring the bridge-wire temperature increase due to a power step excitation. The characterized EEDs were inert (without any explosive material) and empty devices and their caps were cut to introduce the temperature probe. The bridge-wire temperature was measured using a Luxtron surface-type fiber optic thermometry probe attached to the bridge-wire [38]. The power was calculated by measuring the DC current and voltage supplied to the EED. To feed the EED, a variable DC source was connected to its terminals. A plastic support, shown in Fig. 7.1, was used to guide the fiber optic probe to the bridge-wire. Since the bridge-wire diameter is just about 25 μm, the assembly alignment is critical to guarantee that this element is inside the probe’s contact area. Elements position was modified until achieve the highest possible thermal resistance for each case.

![Fig. 7.1. Instrumented EED: Fiber optic probe assembled to an inert EED.](image)

Fig. 7.2 shows the thermal response of an EED. Particularly, this figure shows the bridge-wire temperature increase when the EED is heated during 180 s. The figure shows that, after establishing, the temperature achieves a maximum of 2.55 °C. This value was assumed as the steady state temperature. The figure shows that the measured time constant of this instrumented EED is several times higher than the excepted one of a typical EED, which is between 20 and 60 ms [81]. This difference in the thermal constant is mainly due to two factors: i) heat losses in an inert EED are higher because the bridge-wire is directly in contact with air and ii) the measured temperature
differs from the actual bridge-wire temperature according to the characteristics of the thermal coupling between the bridge-wire and the probe.

Fig. 7.2. Thermal response of an EED excited by a power step.

In [95], it is proposed that the dynamic response of a fiber optic probe attached to a bridge-wire can be described by a first-order low pass filter, as shown in Fig. 7.3. In that circuit, $\theta_{BW}$ and $\theta_M$ are the bridge-wire and the measured temperatures, respectively, $R_{th}$ and $C_{th}$ are the thermal parameters of the EED first-order model, and $R$ and $C$ are the thermal resistance and capacitance representing the probe. For the transient response, the model shows that the measured temperature is defined by the sum of two exponentials; one determined by the EED thermal characteristics and another determined by the probe characteristics. Since the used probe presents a response time of 25 ms [38] and the measured temperature establishes in several tens of seconds, it seems that additional terms should be included in the thermal circuit.

Fig. 7.3. Thermal circuit for the bridge-wire temperature measurement assembly.

Due to the difficulty of having the probe directly in contact with this bridge-wire due to its small diameter, a more general description of the heat flow in the measurement is presented in Fig. 7.4. In this model, the probe is in contact with the material surrounding the bridge-wire. The circuit is
based on the second-order model presented in Chapter 3. In Fig. 7.4, $\theta_1$, $\theta_2$, and $\theta_M$ are the bridge-wire, surrounding materials, and measured temperatures, respectively, $R_i$ and $C_i$, with $i = 1$ and 2, are the thermal parameters of the EED second-order model, and $R$ and $C$ are the probe’s thermal parameters. In this case, the step response presents a third exponential term due to the heat flow to the surroundings. These effects are important if the temperature is the variable of interest and the transient response is required. However, as shown in the next section, the measurement system can be used to quantify the dissipated power in steady-state as a function of the temperature.

![Thermal circuit for a temperature measurement in the bridge-wire surroundings.](image)

**7.2.2 Steady-State Response**

In steady-state, the model of Fig. 7.3 predicts that the measured temperature for a step excitation converges to the bridge-wire temperature. Then, characterizing the slope of the curve of the measured temperature as a function of power would produce the EED thermal resistance. For the model in Fig. 7.4, the steady-state measured temperature is given by:

$$\theta_M = \frac{R_2}{R_1 + R_2}. \quad (3.1)$$

Then, there is a linear deviation that mainly depends on the distance between the probe and the bridge-wire. In this case, these linear effects can be removed calibrating. From Fig. 7.4, the steady-state relation between power and measured temperature yields

$$\theta_M = PR_2. \quad (7.3)$$

Therefore, $R_2$ corresponds to the transfer function between the dissipated power in the instrumented EED and the measured temperature. It can be obtained measuring steady-state temperatures for different power conditions. Fig. 7.5 shows the steady-state temperature as a function of the power for four samples of EEDs. A linear regression was performed for each case and the fit equation is shown in the figures. Notice that the slope of the linear fit corresponds to the
instrumented EED thermal resistance. In the next section, this technique is used to perform a radiation test on EEDs in different configurations.

![Graphs showing steady-state thermal response of EEDs](image)

**Fig. 7.5.** Steady-state thermal response of (a) EED1, (b) EED2, (c) EED3 and (d) EED4.

### 7.3 Radiation Test with a LPM Source

A radiation test of two dipole antennas attached to inert detonators and three samples of inert improvised explosive devices (IED)‡ with inert detonators was performed. The radiated energy was produced by a low power microwave (LPM) generator with output power of 800 W and operating frequency of 2.45 GHz. The power coupled in the IED was measured using a fiber optic thermometer attached to the detonator’s bridge-wire.

#### 7.3.1 Experimental Setup

Fig. 7.6 shows the measurement setup inside the anechoic chamber. The device under test (e.g. the inert IED) was located at 3 m away from the transmitting antenna. The instrumentation and equipments required for the test were:

---

‡ IEDs were provided by Mr. Ernesto Neira from the Escuela de Ingenieros Militares de Colombia
- LPM Radiator: Microwave source (800 W @ 2.45 GHz), waveguide launcher, waveguide, and horn antenna (8 dBi)\(^3\).
- Electric Field Probe Kit - FL7006 Kit
- Industrial Temperature Monitor - Luxtron 812
- Surface Temperature (STS) Fiber Optic Thermometry Probe – Luxtron FOT Probe STS
- Variable Output Voltage Source, 5 V, 2 A.
- 2 Multimeters
- Meter
- Wood support

Fig. 7.6. Experimental setup of the radiation test.

### 7.3.2 Procedure

Table 7.1 summarizes the setup of the performed measurements. First, the magnitude of the incident electric field was measured using the electric field probe. Then, the temperature response of 5 devices was measured. The devices under test (DUTs) were composed by an instrumented EED attached to dipoles (devices # 1 and 2) and attached to inert replicas of IEDs (devices # 4 to 6); some of them are shown in Fig. 7.8. The temperature probe was the same for all measurements. The electric field radiated in measurements 2 to 6 was also measured using the electric field probe, as shown in Fig. 7.7.

In the radiation test of each device, the procedure was:

---

\(^3\) The LPM system was provided by the Electromagnetic Compatibility Group of the Universidad Nacional de Colombia
1. Place the DUT on the tower.
2. Attach the fiber optic probe to the EED’s bridge-wire.
3. Using the temperature monitor, the voltage source and the multimeters, measure the instrumented EED thermal resistance.
4. Configure and activate the LPM radiator.
5. Record the temperature reading.

Table 7.1. Characteristics of the measurements

<table>
<thead>
<tr>
<th>Device #</th>
<th>Device</th>
<th>LPM Output Power (W)</th>
<th>Duration (min)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>Electric Field Probe</td>
<td>800</td>
<td>0.5</td>
</tr>
<tr>
<td>1</td>
<td>Tuned Dipole @ 2.45 GHz + EED1</td>
<td>800</td>
<td>0.5</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>1.5</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>2</td>
</tr>
<tr>
<td>2</td>
<td>7.96 cm Dipole + EED1</td>
<td>800</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>2</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>2</td>
</tr>
<tr>
<td>3</td>
<td>IED5 + EED2</td>
<td>800</td>
<td>2</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>2</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>3</td>
</tr>
<tr>
<td>4</td>
<td>IED11 + EED2</td>
<td>800</td>
<td>2</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>3</td>
</tr>
<tr>
<td>5</td>
<td>ED13 + EED4</td>
<td>800</td>
<td>2</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>3</td>
</tr>
</tbody>
</table>

Fig. 7.7. Experimental setup for measurements number 2 to 6.
7.3.3 Measurements Results

7.3.3.1 Electric Field

The incident field on the DUT, measured using the electric field probe, is presented in Fig. 7.9. The figure shows that the incident field is equal to zero periodically. In fact, this behavior is due to the operating way of the microwave oven magnetron that was used as generator. In these devices, the magnetron is activated only in a half of the AC power cycle. In this case, with a 60 Hz frequency, the magnetron is on during 8.3 ms and off during other 8.3 ms; however, the probe’s sampling rate is just 4 samples/s. For this reason, the measured field is a temporal aliased version. Since a varying signal of the incident electric field was measured, the RMS value of the measured field was calculated. It is compared with the expected incident field in Table 7.2. Although the measured field has aliasing problems, the RMS value is close to the calculated filed using the nominal power of the source.
7.3.3.2 Thermal Resistance Measurement

The thermal resistance was calculated from measurements of steady-state temperatures and DC power supplied to the instrumented EED. As explained earlier, any movement could modify the thermal coupling between the bridge-wire and the probe; however, to reuse the probe, it was not fixed to the EED. For these reasons, these measurements were performed after placing the complete DUT (i.e. instrumented EED + wires) in the test position inside the anechoic chamber.

Fig. 5.1 shows the measurements and the linear regression of the curve Power VS Temperature for DUT# 1. The slope of this linear regression corresponds to the instrumented EED’s thermal resistance \( R_{th} \). The same procedure was used to calculate \( R_{th} \) for the other configurations. Table 7.3 shows the data and the calculated thermal resistance for all the DUTs.

<table>
<thead>
<tr>
<th>Incident Electric Field (V/m)</th>
<th>Difference (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Calculated* ( E )</td>
<td>Measured ( E )</td>
</tr>
<tr>
<td>129.7</td>
<td>149.95</td>
</tr>
</tbody>
</table>

\( |E| = \frac{1}{d} \sqrt{30P_T G_T} \), where \( d = 3 \text{ m} \) is the distance, \( P_T = 800 \text{ W} \) is the MW generator power, and \( G_T = 8 \text{ dB} \) is the antenna gain in linear units.

Fig. 7.9. Electric field measured at the DUT’s distance.

Table 7.2. Measured electric field compared with calculations.

Fig. 7.10. Bridge-wire temperature as a function of the supplied power for DUT # 1.

89
Table 7.3. Data for thermal resistance characterization

<table>
<thead>
<tr>
<th>Device #</th>
<th>Elements</th>
<th>Current (A)</th>
<th>Voltage (V)</th>
<th>Power (W)</th>
<th>Temp. (°C)</th>
<th>Thermal Resist. (°C/W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 &amp; 2</td>
<td>Dipoles + EED2</td>
<td>0.09</td>
<td>0.404</td>
<td>0.03636</td>
<td>26</td>
<td>91.366</td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.25</td>
<td>1.3</td>
<td>0.325</td>
<td>53</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.49</td>
<td>1.4</td>
<td>0.686</td>
<td>85.4</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>IED 5 + EED3</td>
<td>0.001</td>
<td>0</td>
<td>0.000001</td>
<td>23.3</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.1</td>
<td>1</td>
<td>0.01</td>
<td>33.3</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.25</td>
<td>1.3</td>
<td>0.0625</td>
<td>83.3</td>
<td>156.52</td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.5</td>
<td>1.9</td>
<td>0.25</td>
<td>170</td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>IED 11 + EED3</td>
<td>0.1</td>
<td>0.6</td>
<td>0.01</td>
<td>28.7</td>
<td>130.00</td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.25</td>
<td>1.2</td>
<td>0.0625</td>
<td>56.7</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.5</td>
<td>2</td>
<td>0.25</td>
<td>150</td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>IED 13 + EED3</td>
<td>0.1</td>
<td>1.2</td>
<td>0.01</td>
<td>32</td>
<td>84.182</td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.25</td>
<td>2.1</td>
<td>0.0625</td>
<td>73</td>
<td></td>
</tr>
<tr>
<td></td>
<td></td>
<td>0.5</td>
<td>3.2</td>
<td>0.25</td>
<td>158.4</td>
<td></td>
</tr>
</tbody>
</table>

7.3.3.3 Induced Power

The actual functionality of the temperature probe is to measure the dissipated power in the bridge-wire. Table 7.5 presents a comparison between the calculated and measured bridge-wire induced power. Calculated power was obtained using (6.1) and the device impedances listed in Table 7.4. Results show that the difference in magnitude between predicted and measured power is below 2.13 dB. The table also shows the measured incident electric field and other measured parameters of the devices, such as the gain, the transmission coefficient τ, and the thermal resistance $R_{th}$.

Although the parameters used for calculations were previously measured, some of them are still an uncertainly source. The devices’ gains and impedances, for example, could have changed due to manipulation. Since the IEDs wires are not fixed inside the container, their geometry could have changed. In addition, the measured incident field magnitude presents an aliasing problem, as described in earlier.

Table 7.4. Impedances of the components of the DUTs measured at 2.45 GHz

<table>
<thead>
<tr>
<th>Element</th>
<th>Impedance</th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>EED2</td>
<td>Real part (Ω)</td>
<td>40.3</td>
<td>-151.1</td>
</tr>
<tr>
<td>EED3</td>
<td>Real part (Ω)</td>
<td>48.09</td>
<td>-169.3</td>
</tr>
<tr>
<td>Tuned Dipole (5.8 cm)</td>
<td>Real part (Ω)</td>
<td>71.47</td>
<td>-0.72</td>
</tr>
<tr>
<td>Dipole (8.2 cm)</td>
<td>Real part (Ω)</td>
<td>303</td>
<td>346.6</td>
</tr>
<tr>
<td>IED5</td>
<td>Real part (Ω)</td>
<td>224.1</td>
<td>-133</td>
</tr>
<tr>
<td>IED11</td>
<td>Real part (Ω)</td>
<td>767.4</td>
<td>-185.2</td>
</tr>
<tr>
<td>IED13</td>
<td>Real part (Ω)</td>
<td>159.4</td>
<td>62.79</td>
</tr>
</tbody>
</table>
Table 7.5. Comparison between measured and calculated induced power. Other parameters of the characterized devices are also presented.

<table>
<thead>
<tr>
<th>Set</th>
<th>Elements</th>
<th>Gain DUT</th>
<th>τ</th>
<th>El - rms (V/m)</th>
<th>Rth (°C/W)</th>
<th>ΔT (°C) Meas.</th>
<th>Pd (mW) Calculated</th>
<th>Pd (mW) Measured</th>
<th>ΔPd (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Tuned Dipole (5.8 cm) + EED2</td>
<td>1.64</td>
<td>0.32</td>
<td>146.87</td>
<td>91.36</td>
<td>2.60</td>
<td>36.30</td>
<td>28.46</td>
<td>1.06</td>
</tr>
<tr>
<td>2</td>
<td>Dipole (7.96 cm) + EED2</td>
<td>1.80</td>
<td>0.31</td>
<td>143.59</td>
<td>91.36</td>
<td>4.52</td>
<td>36.75</td>
<td>49.47</td>
<td>1.29</td>
</tr>
<tr>
<td>3</td>
<td>IED5 + EED3</td>
<td>1.47</td>
<td>0.26</td>
<td>141.21</td>
<td>156.52</td>
<td>2.55</td>
<td>24.11</td>
<td>16.29</td>
<td>1.70</td>
</tr>
<tr>
<td>4</td>
<td>IED11 + EED3</td>
<td>1.48</td>
<td>0.19</td>
<td>147.25</td>
<td>130.00</td>
<td>1.51</td>
<td>18.95</td>
<td>11.62</td>
<td>2.13</td>
</tr>
<tr>
<td>5</td>
<td>IED13 + EED3</td>
<td>0.34</td>
<td>0.59</td>
<td>144.82</td>
<td>84.18</td>
<td>1.30</td>
<td>13.52</td>
<td>15.44</td>
<td>-0.58</td>
</tr>
</tbody>
</table>

7.4 Conclusions

The steady-state and transient thermal response of EEDs was analyzed. The implementation issues of the EED’s thermal response characterization using fiber optic thermometry measurements were discussed and described using two models. The models included the effect of the probe’s thermal response and the small contact area of the EED’s bridge-wire.

In addition, an electromagnetic susceptibility test of different structures connected to EEDs was performed by measuring the power dissipated in the EED’s bridge-wires. A fiber optic thermometer was used to measure the power to avoid disturbing their electromagnetic behavior. Measurements are in good agreement with results of theoretical models used to calculate induced power in wired EEDs. A difference in the bridge-wire power lower than 2.13 dB was obtained between calculations and measurements.

The electromagnetic field radiated by the LPM source used for the tests was measured. The electric field presented a waveform similar to a pulse train. Unfortunately, the sampling rate of the electric field measurement system was lower than the pulses’ frequency. For this reason, an accurate measurement of this variable was not possible.

The instrumented EEDs presented a slow time response, in the order of tens of seconds. This was attributed mainly to the quality of the thermal coupling between the bridge-wire and the fiber-optic probe. This response could be faster if the heat losses in the coupling are reduced. It could be achieved assuring a better contact with the bridge-wire or including a surrounding medium with low thermal resistance in the junction, such as a gel.
Chapter 8
Summary, Conclusions, and Perspectives

In this thesis, the electromagnetic susceptibility of devices with EEDs is studied analytically, numerically, and experimentally. Special interest is devoted to the device’s response to HPEM disturbances. This thesis proposes procedures, models, and techniques suitable to determine the frequency range in which the susceptibility of an EED connected to a firing circuit with random characteristics increases. The proposed approaches are based in calculating the energy transfer from an incident disturbance to the critical element in the system (i.e. the EED’s bridge-wire) from three transfer functions; which describe the system’s electromagnetic and thermal responses. The electromagnetic response of the EED and its firing circuit is modeled using different strategies based in numerical calculations and analytical approaches. The thermal response, due to the thermal dissipation process inside the EED, is modeled using analytical models reported in the literature, which are verified using numerical simulations. The problem in characterizing the system response due to the variability of the firing circuits is addressed using equivalent structures and Monte-Carlo technique. Additionally, the proposed models are experimentally validated. In the next lines, the main conclusions found in the development of this work and the perspectives of applications of these findings are presented.

General Conclusions

The following main conclusions, detailed next, summarize the general findings of this research:

- The proposed susceptibility model is appropriate to assess the susceptibility of EEDs against different types of EMI. It was shown that EED systems are specially disturbed by CW signals due to their capability to deliver higher energy levels.

- Experimental techniques to characterize the reception properties of devices with EEDs were proposed, implemented, and used to validate the susceptibility model.

- The proposed models showed to be suitable to determine the resonance frequency range and the ignition field intensities of an EED connected to a firing circuit with random characteristics.
Electromagnetic Susceptibility Model of EEDs

In the first chapters, the state-of-the-art of the electromagnetic susceptibility of EEDs is presented. Recent studies on this topic show that EEDs can be initiated by radiated disturbances. In effect, experimental results show that tuned signals, such as damped sinusoidal and CW, can be used to induce currents higher than EEDs all-fire thresholds. These experimental studies indicate that activation with modern IEMI systems is possible. However, a theoretical research on the response of the EEDs was required.

This research intended to study this problem by parts. It was shown that the system’s coupling response, which includes the EED and its firing circuit, can be modeled using three blocks:

\[
\text{External Interaction} \rightarrow \text{Penetration} \rightarrow \text{Port of Interest}
\]

Each block represents the transfer function in different stages in the system. The first transfer function, which corresponds to the External Interaction, relates the incident EM field and the induced currents in the wiring of the firing circuit. Then, these currents are used to calculate, by means of the Penetration transfer function, the power dissipated in the bridge-wire, which is the system’s transducer. The last transfer function, named Port of Interest, relates this power into temperature.

For the external interaction a Norton equivalent circuit is used. It allowed us to calculate independently the firing circuit and the EED responses. It was particularly useful in evaluating the variability of the firing circuit. The Port of Interest was studied using thermal models that describe the bridge-wire power dissipation.

Susceptibility of EEDs against HPEM

Based on numerical simulations, the performance of different thermal models was assessed. It is shown that the thermal models should be selected according to the excitation duration. First-order thermal models described better EED response for periods of time much lower than the EED thermal constant. On the other hand, second-order model worked better for the late-time response. In practice, it means that different models should be used for fast excitations, such a short pulses, that for slow excitations, such as a step function. Practical simplifications for short (or adiabatic) excitations and for steady-stable response are developed.

Using the appropriate model in each block, the response of EEDs in different configurations and using different excitations was calculated. Initially, four HPEM real disturbance sources are assessed on an EED with wires in dipole configuration. From this comparison, which includes
Hypoband, Mesoband and Hyperband disturbances, it was verified that radars are the most perturbing sources due to their higher duty-cycle and tuning possibility. The EED detonation condition was achieved with a ship radar located at 5 m of distance. On the other hand, according to the results with the JOLT system, the radiator of impulse like waveforms with the highest field-range product, very short HPEM pulses (hyperband sources) does not induce enough energy to ignite EEDs at practical distances.

**Experimental Characterization**

Experimental techniques to characterize the reception properties of devices with EEDs were proposed, implemented, and used to validate the susceptibility model.

To validate the electromagnetic model with experimental results, a technique to fully characterize the receiving properties of differential devices was developed. The technique is based on measuring the mixed-mode S parameters to calculate the gain and the differential input impedance. Applying an appropriate calibration and de-embedding processing, this method offers an accurate and a broadband measurement. EEDs and firing circuits were characterized using this technique. As a result, the analytical model used to calculate the EED input impedance was experimentally verified. Another application of this balanced measurement is the characterization of differential antennas. The gain and impedance of simple wire antennas were compared with theoretical calculations showing a good agreement. This technique could also be used in electromagnetic compatibility tests to characterize the EM response of devices with differential ports. In this study, the differential mode parameters were used; however, the common mode information is also available in the mixed-mode S parameters.

To validate the complete model, including the thermal response, a technique to assess the EM susceptibility of EEDs based on fiber optic thermometry was used. This technique allows to perform radiation tests and to obtain the induced power in an EED. The model showed to be in a good agreement with experimental results of a radiation test with a low power microwave source. It was shown that the surface fiber optic sensor can be used for CW excitations; however, it presents a slow dynamic response to be used for transient excitations.

**Susceptibility of EEDs with a random firing circuit**

**Statistical Analysis**

Having in mind that CW disturbances are capable to produce remote detonations, a study on these signals is developed. For these kinds of excitations, the system resonance response is a determinant
factor since it indicates the range of frequencies in which the system receives and dissipates higher power levels. The resonance frequency is a function of the physical characteristics of all the system elements. For this reason, the knowledge of the response variability of each element is required.

The Norton model used to calculate the EM coupling, allows dividing the system in two main components, which are the firing circuit and the EED. Based on a previous study, the EED response was assumed to be deterministic (typical). On the other hand, the firing circuit response was analyzed statistically. The geometry, dimension, elements, enclosure, and surrounding materials of the firing circuit depend on and vary according to the applications and the manipulation. As a consequence, the induced power and the EEDs ignition become a random variable.

A first statistical analysis, representing the firing circuit by a simple variable wiring geometry, was performed applying the Monte-Carlo technique. A representative quantity of wiring realizations was numerically simulated to calculate the random EM response of the induced power as a function of the frequency. This analysis brought two important results that characterize the susceptibility of the wired EED: the system resonance frequency and the induced-power probability density function.

Inert Explosive Device Characterization

A system in which EEDs are used is the Improvised Explosive Device (IED). This was particularly studied with the proposed susceptibility model and experimental technique. As more knowledge of the elements response is included in the model, a better description of the system response is achieved. For this reason, the characterization of IEDs, which have inherent random physical characteristics, was based on a sample of 22 inert IEDs implemented with actual manufacturing techniques. The receiving properties of each sample were measured with the differential technique. In addition, the dimensions, materials, geometries, and firing circuit elements were reported and the samples were classified in three groups according to their switch structure: syringe, “M”, and sponge. Therefore, this characterization not only resulted in the EM response of the IEDs, but also in a better knowledge of the physical properties and elements of IEDs used in Colombia.

This analysis was based on decomposing the EM coupling in two terms: i) the gain and ii) the mismatch between the firing circuit and the EED. They represent, respectively, the ability of receive radiated EM energy and the ability to send to this energy to the port of interest. The gain is affected by the surrounding elements, such as the surrounding medium, the container, and the dielectric materials, and the mismatch is mainly determined by the effective dimensions of the wiring. Each sample presented a particular response depending on the specific characteristics; for example, IEDs
with metallic container presented a frequency selective behavior, in contrast with samples of plastic container. In spite of these particularities, a receiving properties average behavior was determined. Similarly to the preliminary results from the numerical simulations with a simplified wiring, the mean system resonance frequency and the probability density function of the induced power could be determined.

**Numerical and Theoretical Models**

Two models were proposed to replicate the experimental results with IEDs. The proposed susceptibility models showed to be suitable to determine the resonance frequency range and the ignition field intensities of an EED connected to a firing circuit with random characteristics. The first approach was based on Monte-Carlo simulations, using the same simplified structure from the preliminary statistical analysis, but considering the wiring length probability distribution of the measured IEDs. In addition, a theoretical model, based on Mean Effective Gain (MEG) and effective antenna volume concepts, was developed to replicate the mean IED response. Both approaches are in a good agreement with experimental results. From these results, it is concluded that simple structures can be used to model the EM response of wired EEDs; however, they should be based on actual firing circuit configurations to describe the resonance behavior in the microwave range.

The theoretical model presents practical results for special cases. For example, it shows that the MEG converges to the efficiency (i.e. considering the wiring as an antenna) if the angle of incidence is a random variable with uniform distribution. It also shows that the average value of the impedance converges to be real in all the frequency range as the number of considered IEDs increases. In contrast, Monte-Carlo simulations offer the possibility of analyzing better the statistical characteristics of induced power. In fact, the empirical probability distribution functions of the power calculated with Monte-Carlo simulations presented the same behavior of the obtained ones from experimental results. The main difference is that the power levels calculated with the numerical simulation are 4 dB higher in average. This is a consequence of the model simplification, mainly in neglecting the container effect.

More complicated effects of the model simplification can be observed comparing the results of the Monte-Carlo simulations presented in Chapter 4 with the ones presented in Chapter 7. Both models use the same EED, the same wiring geometry, and the same excitation. However, they use different probability distribution function (PDF) and range of the wiring length and PDF of the firing power threshold. These changes modified the system’s resonance frequency (from 600 MHz to 761 MHz), the PDF of the power (from Normal to Extreme Value distribution), and the electric
field to achieve a survivor function of 5% (from 2.5 to 12 kV/m). These changes, even in the PDF, show the importance of the knowledge of the characteristics of the random variables in the device under study.

From these results, it is concluded that the remote activation of IEDs is viable with a source tuned to the mean system’s resonance frequency range and able to radiate high strength EM fields. From the experimental tests, such a source should produce an electric field of 12 kV/m during 2.5 ms on the IED wiring to have 95% of successful remote detonations. As the distance between the source and the IED increases over 30 m, the characteristics of the source become unpractical with the current technologies.

**Future Research**

This study was focussed on the electromagnetic susceptibility of EEDs. Models, techniques, and procedures to characterize the response of EEDs excited by HPEM were developed. They were used to study the response of Improvised Explosive Devices, which is a particular application. However, the proposed models and measurement techniques can be directly applied to study the susceptibility of EEDs in other applications (e.g. mining).

In addition, these techniques can be directly applied for any device in which the Point-of-Interest transfer function is a thermo-electric interaction. That is, devices in which the main effect of the disturbance is heating a critical element. Other electronic systems, for example devices in which the main effect is a semiconductor failure, can also be analysed using the proposed scheme but the Point-of-interest block should be replaced according.

Here, an equivalent structure was used to model a firing circuit with random properties. It is promulgated that this structure replicates the mean response of the samples if it has the average effective height and volume. An interesting research could be focussed in the application of this technique in other devices. A generalization of this technique would be an important contribution in the analysis of the electromagnetic susceptibility.

It is important to note that in the presented models, it was assumed that the EED has a typical (deterministic) EM response. It was based on a previous study in which a representative quantity of EEDs is experimentally characterized and it is observed that the input impedance converges to a typical behavior. However, materials and dimensions used in the EEDs are different according to the manufacturer and the type. The model results, such as the optimal frequency of coupling, are dependent on these characteristics. For this reason, a statistical characterization of the actual EED used in the assessed system should be performed. It can be done using the experimental procedure
or the analytical model discussed in this thesis. If the uncertainty of this parameter is not negligible, the EED input impedance should be included as an input in the statistical analysis. A future work would study the influence of considering this parameter as a random variable in the induced power frequency response. It would be expected that the system resonance frequency results modified.

In the analysis of the Improvised Explosive Devices, some simplifications should be addressed. In this study, inert devices without explosive material were used. In addition, they were placed in a free-space environment, even though in practice, they are partially buried. Further research should address the effects on the coupling due to the electromagnetic properties of the fillers and the soil that surround the firing circuit.

The theoretical model was validated using a radiation test and fiber optic thermometry. These kinds of measurements are an alternative to destructive tests and provide important information of the coupling that can be used in the data analysis. Similar tests, including higher power levels and a frequency sweep in the range of interest, could be used to assess the EM susceptibility of devices in realistic scenarios. Studies to improve the dynamic response of the measurement should be required for transient excitations. Furthermore, the LPM source used here presented an unstable output and a CW pulsed waveform. Work to improve these characteristics could be conducted to achieve a more appropriate reference excitation field.
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List of Publications

Conference Papers


Journal Papers


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John J. Pantoja was born in Pasto, Colombia in 1986. He received the electronic engineering degree from the Universidad Nacional de Colombia, Bogotá, Colombia in 2008. Then, he started the M.Sc. in Electronics and Computing at the Universidad de los Andes, Colombia, where he joined the Group of Electronics and Systems of Telecommunication (GEST). There, he contributed in a project on the detection of unintended electromagnetic emissions of electronic devices. He accomplished his M.Sc. degree in 2013 focusing on the experimental characterization of electric devices susceptibility against high power electromagnetics.

In 2010, he started his Ph.D. program in electrical engineering at the Universidad de los Andes under the supervision of Prof. Néstor Peña. There, he joined the international research project “Application of High Power Electromagnetics to Human Safety” (HPEM-HS), developed by the Swiss Federal Institute of Technology of Lausanne (EPFL), the Universidad Nacional de Colombia, and the Universidad de los Andes. His Ph.D. research involves the numerical modeling and the experimental characterization of the electromagnetic susceptibility of electroexplosive devices and improvised explosive devices. In addition, during his Ph.D. he assisted in measurements of electromagnetic compatibility and antennas performed in the anechoic chamber of the Universidad de los Andes. He received additional financial support from the Swiss Confederation as scholarship holder, between 2011 and 2012, during his internship in the Electromagnetic Compatibility (EMC) Laboratory at the EPFL, directed by Prof. Farhad Rachidi. In addition, in 2013 he received a doctoral fellowship from the Administrative Department of Science, Technology and Innovation from Colombia (Colciencias).

His research areas include electromagnetic compatibility, intentional electromagnetic interference, computational electromagnetics, microwave heating, and social innovation.