

**IMPROVING OVER-THE-AIR ANTENNA PERFORMANCE
TO STABILIZE LTE/5G CELLULAR COMMUNICATION IN
ALTERNATING CURRENT ELECTRIC VEHICLE
CHARGERS**

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by

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To the memory of my dear mother, who inspired me to never give up.



DECLARATION OF ORIGINALITY

I hereby declare that I am the sole author of this thesis and that this is the true copy of my thesis, including the final revisions, approved by my thesis committee. All data and information have been obtained, produced, and presented in accordance with the rules of research ethics and principles of academic honesty. As required by these rules, to the best of my knowledge I have acknowledged ideas, thoughts, and any copyrighted material in accordance with the standard referencing rules. I certify that any part of this thesis has not been submitted for a degree or diploma in another educational institution.

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ABSTRACT

This thesis aims to ensure robust cellular connectivity for alternating current electric vehicle (AC EV) chargers by implementing reliable communication protocols and optimizing antenna performance. Key performance metrics, such as Total Radiated Power (TRP) and Total Isotropic Sensitivity (TIS), are used to assess connectivity between the cellular module, antennas, and the base station. Root-cause analyses identify the factors degrading antenna performance, while theoretical evaluations support the proposed design improvements. The results offer a framework to enhance connectivity, ensuring remote control of AC EV chargers as mandated by the Open Charge Point Protocol. The proposed methodology is validated on two AC EV charger platforms. For the EVC01 platform, low band Long Term Evolution (LTE) degradation is shown to result from near-field interaction between the antenna and Printed Circuit Board (PCB), together with conducted interference components. Restructuring the PCB layout and applying a common mode filter improves LTE Band 8 TIS by approximately 6 dBm. For the EVC08 platform, the dominant issue is near-field coupling induced by the internal panel structure, and suppressing resonance through a dual-point grounding approach improves LTE Band 8 TRP by approximately 6 dBm. Furthermore, for 5G New Radio (NR) Frequency Range 1 (FR1) operation on the EVC08 platform, optimized antenna placement and proper port–band mapping yield approximately 17 dBm improvement in TIS for the n5, n20, and n28 bands, ensuring stable connectivity across all tested channels. This thesis establishes a physically consistent and experimentally validated framework to improve LTE/5G NR FR1 cellular communication performance in AC EV chargers.

Keywords: Total radiated power (TRP), total isotropic sensitivity (TIS), long term evaluation/5G new radio (LTE/5G NR), antenna performance

ÖZET

Bu tez, alternatif akım elektrikli araç şarj cihazlarında güvenilir hücresel bağlantı sağlamak amacıyla iletişim protokollerinin uygulanması ve anten performansının iyileştirilmesine odaklanmaktadır. Bağlantı performansı, toplam ışınlanan güç (Total Radiated Power, TRP) ve toplam izotropik alıcı duyarlılığı (Total Isotropic Sensitivity, TIS) metrikleri üzerinden değerlendirilmiştir. Yapılan kök neden analizleri ve teorik incelemeler, geliştirilen tasarım iyileştirmelerini desteklemiş ve Açık Şarj Noktası Protokolü kapsamında gerekli uzaktan kontrol gereksinimlerini karşılayan bir değerlendirme çerçevesi ortaya koymuştur. Önerilen metodoloji iki farklı şarj platformunda doğrulanmıştır. EVC01 platformunda düşük bant Uzun Süre Evrim haberleşmesi (Long Term Evolution, LTE) performansındaki bozulmanın yakın alan etkileşimi ve iletken girişimden kaynaklandığı gösterilmiş; baskı devre kartının yeniden düzenlenmesi ve ortak-mod filtre uygulanmasıyla LTE Bant 8 TIS değeri yaklaşık 6 dBm iyileştirilmiştir. EVC08 platformunda, panel kaynaklı yakın alan kuplajının çift nokta topraklama ile bastırılması sonucunda LTE Bant 8 TRP değeri yaklaşık 6 dBm artırılmıştır.

Ayrıca EVC08'in 5G Yeni Radyo haberleşmesi (New Radio, NR) Frekans Aralığı 1 (Frequency Range 1, FR1) çalışmasında, anten yerleşiminin optimize edilmesi ve doğru port-bant eşlemesi ile n5, n20 ve n28 bantlarında TIS yaklaşık 17 dBm iyileşmiş ve tüm kanallarda kararlı haberleşme elde edilmiştir. Bu tez, alternatif akım elektrik şarj cihazlarında LTE/5G NR hücresel haberleşme performansının iyileştirilmesine yönelik deneysel olarak doğrulanmış bir çerçeve sunmaktadır.

Anahtar Kelimeler: Toplam ışınlanan güç (TRP), toplam izotropik alıcı duyarlılığı (TIS), LTE/5G NR, anten performansı

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LIST OF ACRONYMS AND ABBREVIATIONS

3GPP	Third Generation Partnership Project
AC EV	Alternating Current Electric Vehicle
ACCT	Control Card
ACPW	Power Card
AFIR	Alternative Fuels Infrastructure Regulation
APAC	Asia Pacific Region
BLER	Block Error Rate
CM	Common Mode
CST	Computer Simulation Technology
CTIA	Cellular Telephone Industries Association
DC	Direct Current
DM	Differential Mode
DRX	Diversity Receive Path
DUT	Device Under Test
DXF	Drawing Exchange Format
EIRP	Effective Isotropically Radiated Power
EMC	Electromagnetic Compatibility

EMI	Electromagnetic Interference
EV	Electric Vehicle
EVC	Electric Vehicle Charger
FPC	Flexible Printed Circuit
FR1	Frequency Range 1
HFSS	High Frequency Structure Simulator
IEEE	Institute of Electrical and Electronics Engineers
IMTC	International Mobile Telecommunications Conference
KCL	Kirchhoff's Current Law
LCD	Liquid Crystal Display
LMHB	Low Mid High Band
LTE	Long Term Evolution
MCU	Microcontroller Unit
MID	Measuring Instruments Directive
MIMO	Multiple Input Multiple Output
NOK	Not Okay / Failed Operation
NR	New Radio
NSA	Non Standalone
OCPP	Open Charge Point Protocol

OK	Okay / Successful Operation
OTA	Over-The-Air
PCB	Printed Circuit Board
PDN	Power Distribution Network
PE	Protective Earth
PRX	Primary Receive Port
PSU	Power Supply Unit
PWM	Pulse Width Modulation
RE	Radiated Emission
RF	Radio Frequency
RX	Receive Path
RFID	Radio Frequency Identification
RGB	Red Green Blue
SMD	Surface Mount Device
SMPS	Switching Mode Power Supply
SoC	System On Chip
SPI	Serial Peripheral Interface
TFT	Thin Film Transistor
TIS	Total Isotropic Sensitivity

TR	Technical Report
TRP	Total Radiated Power
TRX	Transmit Receive Path
TS	Technical Specification
TX	Transmit Path
UART	Universal Asynchronous Receiver/Transmitter
USB	Universal Serial Bus
WCDMA	Wideband Code Division Multiple Access

1. INTRODUCTION

Modern AC EV chargers are no longer simple devices that only supply power; Instead, they operate as systems that rely heavily on communication and are capable of status reporting, fault notification, user authorization, and remote monitoring. The continuity of these functions depends strongly on the stability of the communication established through OCPP between the charger and the central management system. For this reason, the LTE and 5G modem and the antenna are critical components that directly determine the operational integrity of the charging device.

However, the internal geometry of AC EV chargers can create an electromagnetic environment that can significantly limit antenna performance. Large PCB surfaces, cable harnesses, power-electronics circuits, and enclosure geometry may distort the antenna's near-field region and introduce electromagnetic loading that weakens both transmit and receive capability. Initial device-level Over-the-Air (OTA) measurements confirmed this behavior, showing that the charger failed to achieve the expected TRP and TIS levels in several LTE and 5G bands. This observation demonstrates that the stability of cellular communication depends not only on the antenna's intrinsic design but also on the electromagnetic characteristics of the device in which it is integrated.

This thesis aims to understand the electromagnetic mechanisms that limit antenna performance in AC EV chargers, systematically identify the physical interactions that lead to reduced cellular performance, and develop engineering solutions to mitigate these effects.

1.1 Motivation and Problem Statement

The primary communication functions of AC EV chargers include status reporting, fault notification, and the initiation of charging sessions. These functions are executed via OCPP and therefore rely on stable, continuous cellular connectivity. Even minor degradations in antenna performance may cause the device to appear offline, interrupt OCPP messaging, or prevent a charging session from being initiated.

The electromagnetic environment inside the charger is therefore considered a potential contributor to this problem rather than a directly verified cause. PCB surfaces, power converters, switching circuits, and cable structures may introduce capacitive and inductive loading in the antenna's near field, which can reduce radiation efficiency and alter the intended radiation characteristics. Furthermore, electromagnetic noise generated by switching power electronics may degrade receiver sensitivity or lead to link instability, thereby affecting TRP and TIS performance.

Motivated by these considerations, this study systematically investigates the electromagnetic behavior inside AC EV chargers, examines how these mechanisms influence antenna performance, and aims to provide practical design guidelines to minimize connectivity degradation in real world applications.

1.2 Scope of Thesis and Contributions

This thesis investigates the interaction between the antenna and the device structure that influences LTE/5G communication performance in AC EV chargers and proposes practical design improvements to mitigate the resulting performance loss. In this context, the study evaluates the TRP and TIS performance of the charger across multiple frequency bands and examines how surrounding PCB structures, cable routing, and power electronic circuitry affect antenna behavior. In addition, the relationship between internal electromagnetic noise sources and OTA performance is analyzed. Based on these findings,

design modifications are introduced to improve the electromagnetic environment around the antenna and their effectiveness is verified through repeated OTA measurements.

The scope of this thesis is limited to improving the electromagnetic environment of the existing charger design. Field conditions, network parameters, LTE/5G modem firmware behavior, and the development of a new antenna design are intentionally kept outside the scope of this study. The results presented in this thesis provide a practical, measurement-based engineering framework that supports antenna placement decisions, electromagnetic interference (EMI) mitigation, and a better understanding of the relationship between OTA performance and electromagnetic compatibility (EMC) in future AC EV charger development processes.

2. LITERATURE REVIEW

The rapid global adoption of EVs has led to a major transformation in transportation infrastructure. This transformation has not only increased overall energy demand but also necessitated the development of smart and manageable charging infrastructure. EV charging systems are generally classified into two main categories: AC and DC (Direct Current) charging. In AC charging, the on-board charger converts grid energy to DC, and the battery is charged accordingly. Since current and voltage regulation is handled directly by the vehicle, this approach reduces the complexity and cost of the charging station. In DC charging, however, the on-board charger is bypassed, and the AC–DC conversion is performed within the charging station itself. This allows the battery to be supplied with high-power DC, significantly reducing charging time, though at the expense of increased station complexity and cost [1].

In practice, AC and DC charging systems are also classified by rated power and typical charging speed under the European Union’s Alternative Fuels Infrastructure Regulation (AFIR). From this perspective, slow chargers delivering 3.7–7.4 kW nominal power and semi-fast chargers up to 22 kW typically provide AC charging solutions. These devices are generally preferred at homes, workplaces, or locations where vehicles remain parked for extended periods, and are particularly suitable for overnight charging. Fast chargers supplying more than 22 kW are usually DC-output systems; by delivering high-voltage DC, they can bring the battery to about 80% state of charge in approximately 45 minutes. These systems are mostly deployed at professional fast-charging stations and, while significantly shortening charging time, they impose additional load on the grid [2], [3]. The growing number of AC and DC chargers has made remote communication and monitoring of these devices increasingly critical. To meet this need, the OCPP was developed as an open and free protocol that standardizes communication between the charge station and the charge station management system. Initially proposed by the

ElaadNL foundation in the Netherlands, it has been standardized and certified by the Open Charge Alliance since 2014. With its demand response messaging structure, OCPP enables the initiation and termination of charging sessions, fault reporting, and status monitoring to be performed in a uniform way across devices from different manufacturers, thereby ensuring global interoperability [4]. To ensure uninterrupted remote monitoring and control via protocols such as OCPP, it is critical that the cellular modems integrated into charging devices deliver adequate cellular performance. Therefore, their transmission and reception performance must be validated under scenarios that reflect challenging real-world conditions. OTA testing is widely used for this purpose, as it evaluates the end-to-end cellular performance of the device including its antenna and Radio Frequency (RF) front end. The most commonly used OTA metrics are TRP and TIS. TRP represents the total power radiated by the device in all directions, while TIS represents the average isotropic receiver sensitivity. The Cellular Telephone Industries Association (CTIA) OTA Test Plan v3.8.2 specifies the angular steps (15° for TRP, 30° for TIS) used in these measurements and how the results are calculated through spherical integration [5]. The Third Generation Partnership Project (3GPP) Technical Report (TR) 38.810 document defines technical requirements for these metrics, enabling results from different laboratories to be comparable [6]. Since TRP measurements require only a single power reading at each grid point, the process is fast and can typically be completed within minutes. TIS measurements, on the other hand, involve iteratively sweeping input power levels at each angle and determining the sensitivity threshold based on block error rate (BLER) performance. This makes the process significantly longer. It has been reported that TRP measurements on a 15° grid can be completed in under 5 minutes, whereas TIS measurements on a 30° grid with 0.5 dB step size can take over 2 hours [7]. These findings demonstrate that OTA testing, particularly TIS measurements, requires considerable time and resources. TRP and TIS measurements have been extensively studied for mobile phones [8, 9]. This

work fills the gap by applying the same methodology to an industrial product, namely an AC EV charger. Unlike mobile devices, AC EV chargers include high-power converters, large PCB areas, and long cable harnesses. These converters use fast-switching power semiconductors with high dv/dt and di/dt transitions. As a result, they create significant EMI sources within the device [10]. During the antenna performance improvement process, the device's Radiated Emission (RE) test behavior was also taken into consideration. RE tests refer to the unintended electromagnetic energy radiated into free space due to parasitic voltages and currents induced on PCB traces, ground planes, and cable harnesses [11]. Literature reports that high switching speeds and parasitic elements on PCBs can significantly degrade RE test results unless proper layout and filtering techniques are implemented [12].

This study provides demonstrations where design modifications simultaneously reduce RE levels and enhance cellular performance metrics in an industrial AC EV charger. It therefore contributes to the literature from both electromagnetic compatibility and cellular performance perspectives.

3. TECHNICAL PERSPECTIVE

AC EV chargers serve as the interface between the electrical grid and the vehicle's battery. Charging units are typically classified as AC or DC systems. In AC charging, the station transfers grid AC directly to the vehicle, and the onboard charger, located inside the vehicle, handles the AC–DC conversion required for battery charging. This structure makes AC EV chargers suitable for residential, workplace, and semi public environments where longer charging durations are acceptable. DC chargers operate differently. These systems convert AC to DC inside the charging unit and deliver DC power directly to the battery. This approach enables faster charging but requires larger hardware and more complex power electronic stages. DC charging is mentioned here for general context only, as the focus of this thesis is on AC EV Chargers. The AC EV chargers evaluated in this thesis (EVC01 and EVC08) comprise several subsystems, including control electronics, communication hardware, protection components, user-interface elements, and an AC/DC power supply that converts the grid input into a DC bus. This DC bus feeds multiple step down and step up converter stages located on the circuit boards inside the device. The resulting power distribution structure is electrically complex and the electromagnetic activity generated within the enclosure can influence the performance of the LTE/5G communication module. In the following parts of the thesis, the technical details of the charger architecture and the factors influencing antenna behavior are presented to support the OTA performance analysis.

3.1 OCPP-Driven Connectivity Requirements

AC EV chargers rely on uninterrupted network connectivity to support remote control, authorisation, and monitoring functions. Public charging stations are required to provide remote access, and this obligation makes stable connectivity essential for both operational and regulatory reasons. Most commercial systems use the OCPP, which enables

communication between the charger and a cloud based management platform. OCPP communication relies on a stable connection, as interruptions can delay responses, cause incomplete transactions, or prevent the charger from being managed as intended. In AC EV chargers that use LTE or 5G for communication, antenna performance has a direct influence on connection stability. Internal electromagnetic activity, power conversion stages, and the physical placement of the antenna may weaken the signal or reduce receiver sensitivity. These variations can affect the delivery of OCPP messages and the overall responsiveness of the charger. For this reason, maintaining consistent cellular performance is critical for reliable field operation. This thesis examines the impact of the internal design of the AC EV chargers used in this study (EVC01 and EVC08) on LTE and 5G link stability during OCPP communication. The findings form the foundation for the OTA performance improvements presented in the following chapters.

3.2 Cellular Technologies & Regional Band Planning

In this section, the cellular technologies used in the EV chargers and the regional LTE band allocations relevant to their operation are discussed. The aim is to clarify which frequency bands are supported, why these bands are required, and how regional differences affect device deployment and performance.

3.2.1 LTE Bands in the AC EV Charger

The EVC01 and EVC08 chargers utilise LTE networks to maintain a stable connection to the charge management platform. Since LTE spectrum allocations vary from one country to another, products intended for deployment in multiple regions must support more than one frequency band. In this study, the focus is on the LTE bands that are widely used in Europe, although the cellular module inside the chargers is also certified for additional markets. During operation, network conditions can change throughout the day. When this happens, the modem may switch to another LTE band to maintain signal

quality. In areas where 5G coverage is limited or unavailable, the charger primarily communicates over LTE, making stable operation in every supported band crucial. Smooth transitions between bands help prevent short drops in communication during a charging session. Additionally LTE band set evaluated here corresponds to the full LTE capability of the module integrated into the chargers. Variations in frequency and wavelength influence both the radiation efficiency of the antenna and the degree to which nearby electronics and the plastic enclosure interact with the radiated signal. Table 3.1 summarizes the supported LTE bands together with their uplink and downlink operating ranges and commonly associated deployment regions. The band definitions and frequency allocations given in this table follow the specifications of 3GPP Technical Specification (TS) 36.101 [13]. LTE is the primary communication technology used by both chargers, and therefore the behaviour of the device in these bands provides an important basis for interpreting the results presented in the following chapters.

Table 3.1: *LTE bands supported by the AC EV chargers and their operating ranges [13].*

Band	Downlink (MHz)	Uplink (MHz)	Region
B1	2110–2170	1920–1980	Europe / Global
B3	1805–1880	1710–1785	Europe / Global
B7	2620–2690	2500–2570	Europe / Global
B8	925–960	880–915	Europe / APAC
B20	791–821	832–862	Europe
B28	758–803	703–748	Europe / Americas / APAC

3.2.2 5G NR FR1 Bands in the AC EV Charger

The EVC08 charger is equipped with a cellular module that can operate on 5G NR FR1 networks. When 5G coverage is available, the device can leverage the enhanced link quality and capacity offered by these networks. Because spectrum allocations differ from one region to another, products intended for international use must be assessed across the frequency ranges used in the markets where they will be deployed. For this reason,

the evaluation performed in this study includes not only the 5G NR FR1 bands used in Europe but also those deployed in North America and other regions supported by the module. Although the charger supports 5G, its communication still relies heavily on LTE. Many commercial 5G networks continue to operate in the Non Standalone (NSA) mode, where LTE remains part of the radio access structure. In this arrangement, 5G NR FR1 carries user data, while control and signaling procedures depend on LTE. This LTE based control layer, often referred to as the *anchor*, maintains a stable connection when 5G coverage becomes weak or inconsistent. As a result, LTE remains a crucial component of the device's overall communication performance. The 5G NR FR1 bands examined in this study correspond to the full set of frequencies supported by the cellular module used in the EVC08 charger. These bands span different parts of the spectrum, and the resulting changes in wavelength influence both antenna radiation and the way the signal interacts with the enclosure and internal electronics. The supported bands, together with their downlink and uplink operating ranges and regional deployment relevance, are summarised in Table 3.2. The frequency definitions and band allocations listed in this table follow the specifications of 3GPP TS 38.101-1 [14]. Evaluating each band individually provides clearer insight into how the charger behaves under different 5G network conditions.

Table 3.2: 5G NR FR1 bands supported by the EVC08 charger and their operating ranges [14].

NR Band	Downlink (MHz)	Uplink (MHz)	Region
n1	2110–2170	1920–1980	Europe / Global
n3	1805–1880	1710–1785	Europe / Global
n5	869–894	824–849	North America / APAC
n7	2620–2690	2500–2570	Europe / Global
n8	925–960	880–915	Europe / APAC
n20	791–821	832–862	Europe
n28	758–803	703–748	Europe / Americas / APAC
n38	2570–2620	2570–2620	Europe / APAC
n40	2300–2400	2300–2400	APAC / India
n41	2496–2690	2496–2690	Americas / APAC
n66	2110–2200	1710–1780	North America
n71	617–652	663–698	North America
n77	3300–4200	3300–4200	Europe / Global
n78	3300–3800	3300–3800	Europe / Global
n79	4400–5000	4400–5000	APAC

4. FUNDAMENTALS OF THE ELECTROMAGNETIC BEHAVIOR OF ANTENNAS

The antenna used in this thesis is a 40 mm Surface Mount Device (SMD) monopole chip antenna mounted on the PCB of an AC EV charger. Although the physical length of the antenna is fixed, its electromagnetic behavior varies significantly with frequency because the wavelength changes with frequency. As a result, the same antenna can behave as an *electrically small* radiator in the low frequency LTE bands and as an *electrically large* radiator in the higher LTE and 5G NR FR1 bands. This frequency dependent electrical size determines the boundaries of the reactive near field region, the distance at which the Fraunhofer far field begins, and the degree to which the antenna is affected by surrounding structures. In order to correctly interpret the TRP and TIS measurements presented in later chapters, it is necessary to establish the basic definitions of wavelength, electrical size, electromagnetic field regions, and near/far field boundaries.

4.1 Wavelength, Wavenumber, and Electrical Size

In free space, the wavelength λ of an electromagnetic wave is related to the frequency f by

$$\lambda = \frac{c}{f}, \quad (4.1)$$

where c is the speed of light. Equation (4.1) shows that the wavelength decreases as the frequency increases. The corresponding wavenumber k , which describes the spatial phase variation, is defined as

$$k = \frac{2\pi}{\lambda}. \quad (4.2)$$

The *electrical size* of an antenna is expressed as the ratio of its maximum dimension D to the wavelength:

$$\frac{D}{\lambda}. \quad (4.3)$$

Following classical antenna theory, the distinction between electrically small and electrically large antennas is given by

$$D < \frac{\lambda}{2\pi} \quad (4.4)$$

for an *electrically small* antenna, and

$$D > \frac{\lambda}{2\pi} \quad (4.5)$$

for an *electrically large* antenna [15, 16]. These conditions will be used later to classify the 40 mm SMD monopole antenna across LTE and 5G NR FR1 bands.

4.2 Electromagnetic Field Regions Around the Antenna

The electromagnetic fields generated by an antenna are conventionally divided into three regions: the *reactive near field*, the *radiating near field (Fresnel region)*, and the *far field (Fraunhofer region)*. The time average power density is described by the Poynting vector:

$$\vec{S} = \frac{1}{2} \Re\{\vec{E} \times \vec{H}^*\}, \quad (4.6)$$

where $\vec{E}$ and $\vec{H}$ are the electric and magnetic field vectors, respectively. The expression in (4.6) is particularly meaningful when the fields have a predominantly radiative character.

In the Fraunhofer far field, the magnitudes of the electric and magnetic fields are related by the intrinsic impedance of free space:

$$|\vec{E}| = \eta_0 |\vec{H}|, \quad \eta_0 \approx 377 \, \Omega. \quad (4.7)$$

Equation (4.7) holds only in the far field region, where the wave can be approximated as a plane wave. In the near field regions, the local field impedance deviates from η_0 and depends on both distance and direction.

4.2.1 *Reactive Near Field*

In the reactive near field region, the fields are dominated by stored (non radiating) energy. The field components decay rapidly with distance (approximately as $1/r^3$ and $1/r^2$), and the field distribution is strongly influenced by the current and charge distribution on the antenna conductors. In this region, nearby structures such as metal parts, cables, and power electronic components can significantly disturb the antenna fields, making the antenna highly sensitive to its electromagnetic environment.

4.2.2 *Radiating Near Field (Fresnel Region)*

Beyond the reactive near field region, the fields become predominantly radiative, but the wavefront is not yet fully planar. This region is often referred to as the Fresnel region. The antenna radiation pattern begins to form here, but it still depends on the distance from the antenna.

4.2.3 *Far Field (Fraunhofer Region)*

At sufficiently large distances, the fields satisfy the plane-wave relation in (4.7). This region is called the far field region. In this region, the radiation pattern is independent of distance and depends only on the observation angle. Antenna performance metrics such as TRP and TIS are defined and meaningful only under far field conditions.

4.3 Near field and far field Boundaries

For an *electrically small* antenna satisfying (4.4), the radius of the reactive near field region can be approximated by [15]:

$$d_{\text{reactive}} = \frac{\lambda}{2\pi}. \quad (4.8)$$

Equation (4.8) shows that the reactive near field radius scales directly with the wavelength; thus, at low frequencies (large λ) the reactive region becomes larger.

For *electrically large* antennas satisfying (4.5), Keller proposes a more accurate approximation for the reactive near field boundary [16]:

$$d_{\text{reactive}} = 0.62 \sqrt{\frac{D^3}{\lambda}}. \quad (4.9)$$

The expression in (4.9) explicitly shows that the reactive near field radius depends on both the antenna size D and the wavelength λ .

The Fraunhofer far field distance is given by:

$$d_{\text{FF}} = \frac{2D^2}{\lambda}. \quad (4.10)$$

Equation (4.10) defines the minimum distance at which the radiation pattern becomes effectively independent of distance and the plane-wave assumption used in (4.7) holds.

4.4 Electrical Size of the 40 mm SMD Chip Antenna Across LTE and 5G Bands

The maximum physical dimension of the antenna under study is

$$D = 40 \text{ mm}.$$

The LTE and 5G NR FR1 bands considered in this work are:

$$\text{B1, B2, B3, B7, B8, B20, B28, n77, n78, n79}.$$

For each band, the free-space wavelength is calculated using (4.1). In practical units, this can be expressed as

$$\lambda[\text{mm}] = \frac{300000}{f[\text{MHz}]}, \quad (4.11)$$

which is directly derived from (4.1). The electrical size of the antenna is evaluated using the criterion in (4.4) and (4.5). If $D < \lambda/2\pi$, the antenna is classified as electrically small; otherwise it is classified as electrically large. Then, the reactive near field radius

Table 4.1: *Electrical size and reactive near field radius of the 40 mm SMD chip antenna. The wavelength is computed using (4.11); electrical size classification follows (4.4)–(4.5); the reactive near field boundary is obtained using (4.8) or (4.9).*

Band	Frequency (MHz)	λ (mm)	$\lambda/2\pi$ (mm)	Electrical Size	d_{reactive} (mm)
B1	2100	142.9	22.7	Large	13.1
B2	1900	157.9	25.1	Large	12.5
B3	1800	166.7	26.5	Large	12.1
B7	2600	115.4	18.4	Large	14.6
B8	900	333.3	53.1	Small	53.1
B20	800	375.0	59.7	Small	59.7
B28	700	428.6	68.2	Small	68.2
n77	3300	90.9	14.5	Large	16.5
n78	3500	85.7	13.6	Large	16.9
n79	4700	63.8	10.2	Large	19.6

is computed using (4.8) for electrically small cases and (4.9) for electrically large cases. Table 4.1 summarizes the wavelength, electrical size classification, and reactive near field radius for each band.

4.5 Impact of Electrical Size on near field Coupling

The values in Table 4.1 show that, at low frequency LTE bands (B8, B20, B28), the antenna satisfies the electrically small condition in (4.4). In these bands, the reactive near field radius is directly proportional to the wavelength as given by (4.8), resulting in a much larger reactive near field *extent* on the order of 53–68 mm. This implies that the antenna is electromagnetically sensitive over a relatively large volume around the device, making it more strongly affected by nearby structures and EMI sources.

In contrast, at middle and high frequency LTE bands and 5G NR FR1 bands, the antenna fulfills the electrically large condition in (4.5). In this case, the reactive near field boundary is governed by (4.9), leading to a considerably *smaller reactive near field extent* in the range of 12–20 mm. As a consequence, the antenna becomes less sensitive to objects located beyond this reduced reactive near field region, and its radiation behavior is more stable with respect to environmental variations. This variation in electrical size

and in the extent of the reactive near field with frequency provides the physical basis for the antenna performance differences across frequency bands that will be discussed in the subsequent chapters.



5. OTA ACTIVE ANTENNA TESTS

In LTE cellular communication systems, the actual antenna performance of a device is determined not only by its isolated passive characteristics but by the complete electromagnetic environment in which it operates. When embedded inside a product, the antenna interacts with power converters, digital circuitry, PCB return currents, flexible printed circuit (FPC) cables, and the plastic or composite enclosure. These interactions significantly alter the reactive near field distribution around the antenna and can cause performance degradations that do not appear in passive measurements [17, 18]. For this reason, evaluating the true radiated performance of a cellular device requires OTA active testing, carried out while the device operates with its own LTE modem and RF front-end [5, 6]. OTA testing considers the antenna as an integral part of the complete device, rather than as an isolated radiating element. With no conducted RF connections, the device transmits and receives entirely through its embedded antenna, allowing realistic characterization of both radiation and sensitivity performance under operational conditions. In LTE systems, OTA measurements yield two primary system level metrics: TRP, representing uplink capability, and TIS, representing downlink reception quality.

5.1 Difference Between Passive and Active Antenna Tests

In passive antenna tests, the antenna is examined independently through parameters such as input reflection coefficient, efficiency, and passive gain. While these parameters provide valuable baseline information, they fail to represent the antenna's behavior when surrounded by real device structures. Power electronics, switching regulators, grounding networks, cable harnesses, and the device enclosure can modify the antenna's near field region and degrade its effective gain pattern [19]. Consequently, an antenna that performs well under passive conditions may exhibit significantly different behavior once integrated into the final product. Active OTA testing reveals such discrepancies by

evaluating the antenna together with the complete device operating in its intended configuration. As emphasized in the classical perspective presented in the *Antenna Engineering Handbook* [20], an antenna cannot be assessed independently from the structure in which it operates; instead, it must always be treated as part of a larger electromagnetic system. TRP and TIS serve as reliable system indicators that capture the combined influence of the antenna, device architecture, surrounding materials, and internal noise sources [21].

5.2 Need for Active OTA Testing in LTE Systems

In practical LTE deployments, antenna performance is strongly dependent on the holistic electromagnetic environment surrounding it. Devices such as EV chargers, wireless gateways, and IoT terminals often include switching converters, large plastic housings, high-current wiring, and wide PCB structures, all of which can distort the antenna's reactive and radiating near fields. These effects are not observable in passive measurements, since the antenna is isolated from the device environment during such tests. Active OTA measurements capture these interactions directly and provide an accurate representation of how the device communicates with the base station under real operational conditions. TRP and TIS thus constitute the most reliable metrics for evaluating LTE system performance. TRP quantifies the overall uplink radiated power capability of the device, while TIS characterizes its downlink reception sensitivity. Together, they form the foundation for the assessment of OTA performance for modern cellular devices.

5.3 Electromagnetic Foundation of OTA Measurements

The theoretical basis of OTA measurements lies in the classical directional radiation behavior of antennas. As described extensively in the antenna literature [15, 18, 19], an antenna does not radiate uniformly in all directions; instead, the radiated power varies with the angular direction and is described by the gain function $G(\theta, \phi)$. In the far field

region, the radiated power flux density in a given direction is denoted by $S(\theta, \phi)$, and the corresponding radiation intensity is defined as:

$$U(\theta, \phi) = r^2 S(\theta, \phi), \quad (5.1)$$

representing the radiated power per unit solid angle. Using (5.1), the TRP of an antenna is expressed as:

$$P_{\text{rad}} = \int_{4\pi} U(\theta, \phi) d\Omega. \quad (5.2)$$

Equation (5.2) establishes the physical foundation of the TRP definition used in OTA testing. Since antenna gain and radiation efficiency are affected by device materials, mechanical layout, and internal EMI mechanisms, the directional radiated power captured in OTA measurements reflects the actual behaviour of the integrated antenna within the device [17, 22].

5.4 OTA Measurement Geometry and Angular Scanning

OTA active measurements are performed in a spherical coordinate system using the angular variables θ and ϕ . The measurement probe samples the radiated power around the device for polar angles $0^\circ \leq \theta \leq 180^\circ$ and azimuth angles $0^\circ \leq \phi \leq 360^\circ$, producing a complete three-dimensional radiation map. This full angular scan provides the directional distribution required to compute TRP in the next section. The measured angular dataset constitutes the foundation for computing the TRP, which is obtained by averaging the radiated power contributions from all directions.

5.5 Total Radiated Power

TRP is the fundamental metric used to quantify the uplink radiated performance of LTE/5G devices under OTA conditions. The TRP represents the *total average power*

radiated by the device in all directions and therefore captures the entire system level behavior of the integrated antenna, RF front-end, enclosure, and internal electromagnetic environment [5, 6].

In antenna theory, the radiated power in a given direction is characterized by the effective isotropically radiated power (EIRP), which accounts for both the transmit power of the modem and the directional gain of the antenna [15, 17]. Consequently, TRP is defined as the spherical average of the directional EIRP distribution:

$$TRP = \frac{1}{4\pi} \int_{4\pi} EIRP(\theta, \phi) d\Omega. \quad (5.3)$$

In practical OTA measurements, (5.3) is evaluated using discrete angular samples collected by the chamber. Thus, TRP is computed as

$$TRP \approx \frac{1}{N} \sum_{i=1}^N EIRP_i, \quad (5.4)$$

where N denotes the total number of angular sampling points. This expression forms the core computational method used by modern OTA chambers. Although this formulation explains how the TRP is numerically obtained, its practical relevance becomes clearer when the physical implications of the metric are considered. Since $EIRP(\theta, \phi)$ is directly influenced by antenna radiation pattern, PCB currents, nearby dielectric and metallic structures, mechanical enclosure geometry, switching converters, cable assemblies and internal EMI sources [19, 20, 22], the actual radiation pattern of an integrated device often deviates from its ideal free space behavior. These effects introduce angular distortions, local nulls, and power suppression regions, which reduce the spherical integral in (5.3) and consequently lower the TRP [20].

A key implication of (5.3) is that strong radiation confined to a limited angular region does not guaranty high uplink performance. TRP penalizes spatial non-uniformity

by averaging radiation over the entire 4π steradian domain. Therefore, rather than reflecting only peak radiation, TRP provides a global and statistically meaningful representation of the uplink capability toward base stations in arbitrary directions.

In practical deployments, these characteristics translate directly into real product performance. For AC EV chargers, the combined effect of antenna placement, PCB layout, mechanical structure, and internal cabling determines the resulting radiation pattern observed in the OTA chamber. Insufficient radiated power in certain angular regions can lead to reduced uplink robustness, intermittent OCPP connectivity, and lower service reliability. Consequently, TRP is widely recognized as a reliable OTA performance metric for LTE/5G devices, as it captures the combined influence of the antenna, device integration, enclosure, and internal electromagnetic behavior.

5.6 Total Isotropic Sensitivity

TIS is the primary OTA metric used to evaluate the downlink reception capability of LTE devices. While TRP quantifies the radiated transmit power of the device in the uplink direction, TIS represents the minimum received power level required to maintain a stable downlink connection. As such, TIS reflects the combined effects of antenna gain, device-level electromagnetic noise, and integration related variations that influence the directional reception performance of the system.

According to the fundamental link-budget formulation, the power received by the antenna in a given direction is expressed as

$$P_{RX}(\theta, \phi) = P_{inc} \cdot G(\theta, \phi), \quad (5.5)$$

where P_{inc} denotes the incident field strength generated by the OTA chamber probe, and $G(\theta, \phi)$ is the directional gain of the integrated antenna. A downlink link can be sustained only if the received power exceeds the sensitivity threshold P_{sens} of the LTE receiver. TIS

is therefore defined as the spherical average of the minimum required sensitivity power over all angular directions:

$$TIS = \frac{1}{4\pi} \int_{4\pi} P_{\text{sens}}(\theta, \phi) d\Omega. \quad (5.6)$$

In practical OTA measurements, (5.6) is computed using discrete angular samples collected by the measurement chamber. The numerical approximation of TIS is then given by

$$TIS \approx \frac{1}{N} \sum_{i=1}^N P_{\text{sens},i}, \quad (5.7)$$

where $P_{\text{sens},i}$ is the measured sensitivity threshold at the i th angular position and N is the total number of sampling points. The physical meaning of TIS is directly tied to the directional receiving performance of the antenna. Variations in antenna gain, the mechanical layout of the device, PCB current distributions, and electromagnetic interference generated by power electronic circuitry can degrade the reception capability in certain angular regions. These weak reception regions increase the global average sensitivity represented by the TIS value. Therefore, a lower (i.e., better) TIS value indicates that the device maintains sufficient downlink reception performance across all directions.

Unlike TRP, which is mainly influenced by transmitted power and radiation efficiency, TIS is sensitive to both the antenna gain and the internal electromagnetic environment of the device. Power converters, cable harnesses, ground structures, and enclosure geometry can introduce direction-dependent noise or coupling that raises the sensitivity threshold. For this reason, antenna placement, PCB design, grounding strategy, cable routing, and EMI management all play a critical role in optimizing TIS performance.

TIS becomes particularly important for stationary devices such as EV chargers. Unlike mobile devices whose orientation changes during use, EV chargers operate in a fixed physical position. As a result, any directional weakness in receiving performance

remains permanent and can directly limit downlink connectivity. Poor TIS may lead to reduced stability in base-station communication, difficulties during session establishment, or delays in OCPP message exchange.

In summary, TIS is a comprehensive metric for evaluating the reception robustness of downlink LTE devices. Reflects directional antenna gain variations, device-induced electromagnetic noise, mechanical integration effects, and receiver sensitivity under realistic operating conditions. For fixed LTE installations, such as EV chargers, TIS is a critical indicator of reliable and consistent network performance.



6. ELECTROMAGNETIC NOISE SOURCES

Electromagnetic noise is an inherent and often critical aspect of modern electronic systems. Even though it is not intentionally radiated, it can couple into sensitive components, distort signals, and degrade system performance. This chapter discusses the physical origins of such noise and provides a framework for understanding how it is generated, how it propagates, and how it influences practical electronic designs.

6.1 Fundamentals of high frequency Electromagnetic Behavior

In this chapter, electromagnetic behavior is considered from an EMI perspective, where the focus is on unwanted electric and magnetic field interactions between circuit structures rather than on intentional radiation from antennas. The electromagnetic behavior of any electrical circuit is fundamentally governed by the Maxwell equations. However, since these equations describe electromagnetic fields as functions of both space and time, their direct solution is impractical for most practical engineering applications. For this reason, electromagnetic interactions are commonly analyzed using circuit models under clearly defined assumptions. The primary assumption underlying such analyses is that the physical dimensions of the structure under consideration are electrically small compared with the signal wavelength. Under this condition, voltages and currents along conductors can be assumed to be approximately uniform, allowing electromagnetic interactions to be represented using equivalent lumped circuit elements. According to the (4.1) as the operating frequency increases, the wavelength decreases, and the electrical size of the structure increases. The lumped-element approximation remains valid when the characteristic physical dimension of the structure, denoted by L , satisfies

$$L \ll \frac{\lambda}{10}. \quad (6.1)$$

When this condition is fulfilled, electromagnetic interactions can be treated as quasi-static, meaning that wave propagation effects can be neglected and the fields are dominated by local energy storage rather than radiation. In this regime, electric field interactions can be represented by equivalent capacitances, while magnetic field interactions can be represented by equivalent inductances. These representations do not imply the absence of electromagnetic fields, but rather that their spatial variation is negligible within the structure.

As frequency increases and the physical dimensions of the structure become a significant fraction of the wavelength, the assumptions underlying lumped-element models gradually lose validity. In such cases, distributed behaviour and field propagation effects must be taken into account, and antenna related phenomena may become relevant. Therefore, the interpretation of electromagnetic noise coupling mechanisms must always be carried out within the validity limits of the adopted modelling approach.

6.2 Scope and Relation to Antenna near field Sensitivity

In Chapter 4, the electrical size of the antenna and its associated electromagnetic field regions were analysed to explain the frequency-dependent sensitivity of the antenna to its surrounding environment. That analysis demonstrated that, at low frequency LTE bands, the antenna operates with an extended reactive near field region. Consequently, the antenna becomes highly sensitive to nearby conductive structures and local electromagnetic disturbances inside the device.

Building upon this observation, the objective of the present chapter is to identify and analyse the physical mechanisms through which electromagnetic noise generated within the device interacts with the antenna's near field region. Rather than focusing on antenna radiation mechanisms, this chapter addresses how unintended electromagnetic noise is generated, coupled, and propagated within the system, ultimately degrading the

measured OTA performance metrics such as TRP and TIS.

From an EMC perspective, electromagnetic noise does not reach sensitive circuits directly. Instead, it propagates along a structured path consisting of a noise source, a coupling channel, and a receptor. This conceptual framework, commonly adopted in EMC analysis, provides a systematic means to understand how local noise sources inside complex electronic systems affect RF performance. The general noise propagation model is shown in Figure 6.1 [23].



Figure 6.1: *Conceptual noise propagation model illustrating the structured path of electromagnetic interference from the noise source to the receptor through a coupling channel [23].*

Within this framework, the antenna is treated as the final receptor, while the subsequent sections of this chapter focus on identifying the dominant noise sources and near field coupling mechanisms present inside the device. By explicitly linking near field electromagnetic coupling mechanisms to antenna near field sensitivity, this chapter establishes the physical basis for the RF desense phenomena observed in the experimental results. Although the term near field is used both in antenna theory and in EMC analysis, it refers to different physical interpretations depending on the reference. In antenna theory, the near field is defined with respect to the antenna itself, whereas in EMC analysis it is defined with respect to the electromagnetic noise source. To avoid ambiguity, the key differences between antenna near field and EMI noise near field concepts are summarised in Table 6.1 [15].

6.3 Noise Sources in high frequency Electronic Systems

In high frequency electronic systems, electromagnetic noise originates from time varying voltages and currents associated with normal circuit operation. Any circuit node

Table 6.1: *Conceptual comparison between antenna near field and EMI near field interpretations, based on classical antenna theory and EMC coupling models[15].*

Aspect	Antenna near field	EMI near field
Reference	Antenna	Noise source
Primary purpose	Antenna operation	Unintended coupling
Field origin	Fields generated by the antenna	Fields generated by the noise source
Distance criterion	Antenna size / λ	Source size / λ
Analysis framework	Antenna theory	EMC and coupling analysis

that exhibits rapid voltage transitions or draws non-uniform current can act as a noise source capable of exciting surrounding electromagnetic fields. According to the framework introduced in the previous section, such noise sources represent the starting point of the noise propagation path shown in Figure 6.1.

Typical noise sources in embedded electronic systems include switching power converters, digital clock signals, high speed data interfaces, and return path discontinuities. These sources are characterized by steep voltage or current edges, which contain significant high frequency spectral components even when the fundamental operating frequency is relatively low. From a physical standpoint, a time varying voltage at a conductor produces a time varying electric field in the surrounding space, while a time varying current generates a corresponding magnetic field. In the lumped-element regime discussed in Section 6.1, these field interactions can be equivalently represented by parasitic capacitances and inductances.

Depending on the electrical characteristics of the noise source, the resulting near field interaction may be dominated either by electric field or magnetic field components. In EMC analysis, these interactions are commonly classified as capacitive or inductive near field coupling mechanisms, which can be described using equivalent circuit representations in the lumped-element regime. A summary of the fundamental differences between these near field coupling mechanisms is provided in Table 6.2 [15].

This comparison highlights that the near field concept defined in antenna theory and the near field approach used in EMI analysis are based on different physical objectives

Table 6.2: *Conceptual comparison between antenna near field and EMI near field interpretations, based on classical antenna theory and electromagnetic coupling models*

Aspect	Antenna near field	EMI near field
Reference structure	Radiating antenna element	Unintended noise source or conductor
Primary purpose	Intentional electromagnetic radiation	Unintentional electromagnetic coupling and interference
Dominant field components	Electric and magnetic fields present simultaneously	Predominantly electric or magnetic field, depending on the source
Physical interpretation	Reactive field behavior associated with radiation mechanisms	near field coupling mechanism between nearby structures
Equivalent modeling approach	Antenna field equations	Lumped-element capacitive or inductive coupling models
Distance criterion	Defined with respect to antenna size and wavelength λ	Defined with respect to noise-source dimensions and wavelength λ
Analysis framework	Classical antenna theory	Electromagnetic compatibility analysis
Typical application	Radiation and pattern characterization	EMI analysis and interference mitigation

and evaluation frameworks. In this study, the EMI oriented near field definition is adopted for the analysis of unintended electromagnetic interactions.

6.4 Capacitive (electric field) Coupling

Capacitive coupling, also referred to as electric field (E-field) coupling, is one of the dominant near field electromagnetic interference mechanisms in electrically small structures. This mechanism arises when a time varying voltage on one conductor produces an electric field that extends into the surrounding space and induces a displacement current through parasitic capacitances into a nearby structure.

Unlike radiated coupling, capacitive coupling does not involve the formation of propagating electromagnetic waves. Instead, the interaction remains confined to the near field region and can be accurately represented using lumped circuit elements, provided that the physical dimensions of the structure are small compared with the wavelength, as discussed in Section 6.1

6.4.1 Physical Origin of Capacitive Coupling

According to Maxwell's equations, a time varying electric field generates a displacement current even in the absence of a conductive path. This displacement current represents the fundamental physical origin of capacitive coupling. From a circuit-theoretic perspective, the electric field interaction between two nearby conductors can be modeled as a parasitic capacitance. When the lumped-element condition is satisfied, this representation enables near field electric coupling to be analyzed using conventional circuit analysis techniques. A rapidly changing voltage on an aggressor conductor generates an electric field whose strength is proportional to the time derivative of the voltage. This field induces a displacement current that flows through the parasitic capacitance toward the victim structure, where it is converted into an undesired noise voltage across the impedance connected to that structure.

6.4.2 Capacitive Coupling Model and Equivalent Circuit

The classical capacitive coupling model proposed by Ott is shown in Figure 6.2 [23]. The figure presents both the physical representation of two nearby conductors and

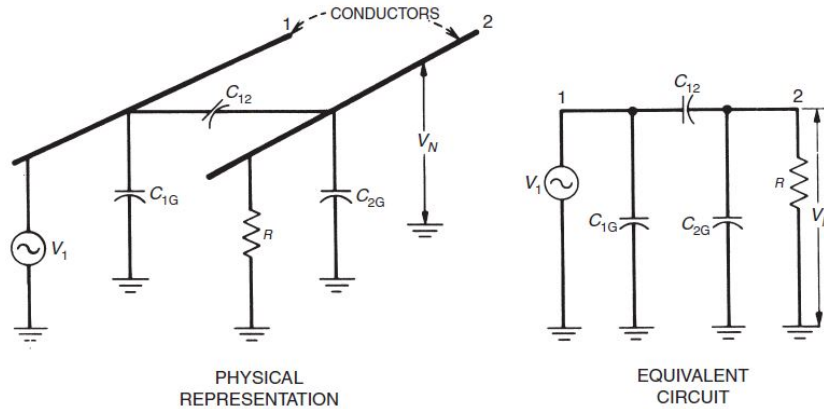


Figure 6.2: Physical representation and lumped-element equivalent circuit of capacitive coupling between two conductors, adapted from the classical model described by Ott [23].

the corresponding lumped-element equivalent circuit used for analytical evaluation.

In this model, Conductor 1 represents the aggressor structure driven by a noise voltage source V_1 , while Conductor 2 represents the victim structure on which an undesired noise voltage V_N is induced.

The parasitic elements are defined as follows:

- C_{12} : mutual (stray) capacitance between the aggressor and victim conductors,
- C_{2G} : total parasitic capacitance between the victim conductor and the reference ground,
- R : equivalent impedance connecting the victim conductor to ground, representing grounding quality, shield termination, or load impedance.

The objective of the analysis is to determine the induced noise voltage V_N on the victim conductor as a function of frequency, parasitic capacitances, and the grounding impedance.

6.4.3 Node-Voltage (KCL) Analysis of the Equivalent Circuit

The equivalent circuit in Figure 6.2 is analyzed in the frequency domain using nodal analysis. The voltage at the aggressor node is fixed by the ideal source and is given by

$$V(1) = V_1. \quad (6.2)$$

The voltage at the victim node is

$$V(2) = V_N, \quad (6.3)$$

which is the quantity of interest.

At node 2, three current paths exist: (i) through the mutual capacitance C_{12} toward node 1, (ii) through the capacitance C_{2G} toward ground, and (iii) through the resistance R toward ground.

In the frequency domain, the admittance of a capacitor is given by $j\omega C$. Applying Kirchhoff's Current Law (KCL) at node 2 yields

$$I_{12} + I_{2G} + I_R = 0. \quad (6.4)$$

The individual current components are expressed as

$$I_{12} = j\omega C_{12}(V_N - V_1), \quad (6.5)$$

$$I_{2G} = j\omega C_{2G}V_N, \quad (6.6)$$

$$I_R = \frac{V_N}{R}. \quad (6.7)$$

Substituting these expressions into the KCL equation gives

$$j\omega C_{12}(V_N - V_1) + j\omega C_{2G}V_N + \frac{V_N}{R} = 0. \quad (6.8)$$

Rearranging terms leads to

$$V_N \left[j\omega(C_{12} + C_{2G}) + \frac{1}{R} \right] = j\omega C_{12}V_1. \quad (6.9)$$

Solving for the induced noise voltage yields

$$V_N = \frac{j\omega C_{12}}{j\omega(C_{12} + C_{2G}) + \frac{1}{R}} V_1. \quad (6.10)$$

Equation (6.10) corresponds to the general capacitive coupling expression derived by Ott [23].

6.4.4 Effect of Victim Impedance: Floating and Grounded Structures

The magnitude of the induced noise voltage strongly depends on the relative magnitude of the grounding impedance R compared with the capacitive impedance. Two limiting cases are of particular practical relevance.

6.4.4.1 high impedance (Floating) Victim Structure

When the victim structure is floating or weakly grounded, the grounding impedance is sufficiently large, satisfying the following condition:

$$\frac{1}{R} \ll \omega (C_{12} + C_{2G}). \quad (6.11)$$

Under the condition stated in (9.3), the resistive admittance is negligible compared with the capacitive admittance, and the induced noise voltage is dominated by the capacitive voltage division between C_{12} and C_{2G} . Consequently, even a small displacement current can generate a relatively large noise voltage on the victim node. Substituting the condition given in (9.3) into the general capacitive coupling expression, the induced noise voltage reduces to

$$V_N \approx \frac{j\omega C_{12}}{j\omega (C_{12} + C_{2G})} V_1 = \left(\frac{C_{12}}{C_{12} + C_{2G}} \right) V_1. \quad (6.12)$$

Equation (6.12) indicates that, in the high impedance (floating) case, the induced noise voltage is independent of frequency and is determined solely by the ratio of the mutual capacitance C_{12} to the total capacitance connected to the victim structure. Physically, this corresponds to a capacitive voltage divider formed between the aggressor–victim coupling capacitance and the stray capacitance from the victim structure to ground.

Because the victim structure lacks a low impedance path to the reference ground, the induced displacement current cannot be effectively discharged and instead accumulates as electric charge on the structure. As a result, the victim behaves as an RF-floating

conductor, making it highly susceptible to capacitive noise coupling even when the displacement current magnitude is relatively small.

6.4.4.2 Low impedance (Well Grounded) Victim Structure

When the victim structure is well grounded, the grounding impedance is small such that

$$R \ll \frac{1}{\omega(C_{12} + C_{2G})}. \quad (6.13)$$

In this case, the resistive admittance dominates the denominator of (6.10), and the induced noise voltage simplifies to

$$V_N \approx j\omega RC_{12}V_1. \quad (6.14)$$

This result indicates that the displacement current preferentially flows through the low impedance grounding path rather than charging the victim structure, leading to a substantial reduction in the induced noise voltage. This behavior explains why effective grounding and proper shield termination are fundamental techniques for mitigating capacitive coupling.

6.4.5 Significance in near field Noise Coupling

Capacitive coupling is inherently a near field phenomenon. In this regime, electric fields do not propagate as radiated waves but remain confined to the immediate vicinity of the conductors. As a result, coupling strength is governed primarily by geometric spacing, parasitic capacitances, and impedance relationships rather than far field radiation mechanisms.

In compact electronic systems, capacitive coupling provides an efficient mechanism through which internally generated noise can interact with sensitive structures before

radiated effects become dominant. Although capacitive coupling is commonly discussed within the EMC literature, its impact on internal electromagnetic conditions directly influences RF performance by modifying local impedance environments and increasing the effective noise level at sensitive nodes.

6.5 Inductive (magnetic field) Coupling

Inductive coupling, also referred to as magnetic field (H-field) coupling, arises from time varying currents that generate magnetic fields linking nearby conductive loops. Unlike capacitive coupling, which is driven by voltage variations and electric fields, inductive coupling is governed by current variations and magnetic flux.

In the near field regime, a changing current $i(t)$ flowing through an aggressor conductor produces a magnetic field whose strength is proportional to the current magnitude and rate of change. If a nearby victim structure forms a conductive loop, part of this magnetic flux links the loop and induces a voltage according to Faraday's law of electromagnetic induction.

From a circuit perspective, inductive coupling can be represented using a mutual inductance M between the aggressor and victim loops. The induced noise voltage on the victim structure can be expressed as

$$V_N = j\omega MI, \quad (6.15)$$

where I is the aggressor current and $\omega = 2\pi f$ is the angular frequency. This relationship shows that inductive coupling strength increases with frequency, current amplitude, and the degree of magnetic flux linkage between the two loops. In practical electronic systems, inductive coupling becomes significant when large current loops exist, such as those formed by switching power converters or poorly controlled return paths. The dominant

mitigation strategies therefore include minimizing loop areas, ensuring tight coupling between forward and return current paths, and reducing high frequency current content.

While capacitive and inductive coupling mechanisms describe how electromagnetic fields are generated and interact with nearby structures in the near field region, they do not fully explain how the resulting noise affects the system as a whole. The magnitude and overall impact of these fields are strongly dependent on the return paths and structural characteristics of the underlying noise currents. For this reason, electromagnetic noise must also be examined in terms of Differential Mode (DM) and Common Mode (CM) current components, which provide a current based classification of the dominant noise propagation mechanisms. To clarify the overall conceptual flow of Chapter 6, the hierarchical relationship between the electromagnetic noise mechanisms discussed throughout this chapter is summarized in Table 6.3 [23, 24].

Table 6.3: *Conceptual classification of electromagnetic noise mechanisms addressed in Chapter 6.*

Key Question	Addressed Concept	Chapter 6 Section
How is electromagnetic noise generated as a field?	Capacitive and inductive coupling mechanisms	Sections 6.4–6.5
Which current structures generate these fields?	Differential mode and CM currents	Sections 6.6–6.7
How far do the resulting electromagnetic fields propagate?	Radiated and conducted noise mechanisms	Sections 6.8–6.9
Why do these mechanisms degrade antenna performance?	RF desense and sensitivity degradation	Section 6.10

6.6 Differential Mode Radiation

Electromagnetic radiation from digital circuits can be decomposed into DM and CM components. DM radiation is associated with the normal operation of the circuit and arises from signal currents flowing in closed loops formed by the signal and its return conductor, as shown in Figure 6.3. These current loops behave as small loop antennas that mainly radiate magnetic fields. Although such loops are required for the functionality of the circuit, their enclosed area and the magnitude of the circulating current must be

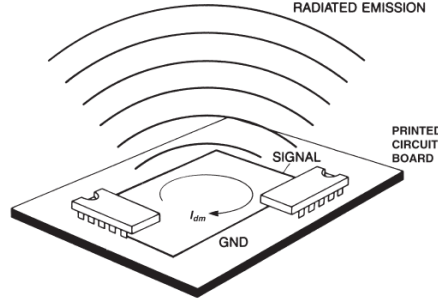


Figure 6.3: *DM radiation mechanism modeled as a small loop antenna, where the signal and its return current form a loop that predominantly radiates magnetic fields [16]*

minimized during the PCB layout in order to reduce the resulting electromagnetic radiation [23].

For a small loop of area A carrying a DM current I_{dm} , the far field electric field magnitude E at a distance r can be approximated as

$$E \propto \frac{f^2 A I_{dm}}{r}, \quad (6.16)$$

where f is the frequency. Equation (6.16) shows that the radiated field increases with the square of the frequency, with the loop area, and with the DM current. Consequently, DM radiation can be controlled by (i) reducing the loop area, (ii) reducing the high frequency content of the current, or (iii) lowering the current amplitude [23].

6.7 Common Mode Radiation

CM currents are defined as currents that flow in the same direction on multiple conductors with respect to a common reference, such as the device chassis or the protective earth (PE) conductor. Unlike DM currents, which form a closed loop between a signal and its return path, CM currents return through parasitic paths including the chassis, cable shields, the PE conductor, or capacitive coupling to nearby structures. As a result, CM currents propagate along larger physical structures compared to DM currents and therefore generate significantly stronger electromagnetic radiation.

In AC EV chargers, long AC input cables, the PE conductor, cable harnesses, shielded communication cables, and extended metal surfaces act as electrically long conductors that behave as unintended dipole like radiators when driven by CM currents. The fundamental physical model for CM radiation is expressed as

$$E_{cm} \propto \frac{f L I_{cm}}{r}, \quad (6.17)$$

where f is the frequency, L is the effective conductor length participating in CM radiation, I_{cm} is the CM current amplitude, and r is the distance to the observation point. Equation 6.17 shows that the radiated field increases with both frequency and conductor length, explaining why physically large structures inside EV chargers produce substantial CM emissions.

CM radiation is most dominant in the 30MHz–300MHz range, where cable lengths become comparable to a significant fraction of the wavelength. However, RE limits are evaluated over the wider 30MHz–1GHz band in EMC standards, causing CM currents to be the primary contributors to measured REs across the full regulatory frequency range.

6.8 Radiated Noise (30 MHz–1 GHz)

Radiated EMI in the 30MHz–1GHz band is primarily driven by the CM mechanism described in Equation 6.17. Long cable structures, the PE conductor, and the chassis behave as efficient radiators when excited by CM currents, resulting in strong electric and magnetic fields in the vicinity of the antenna.

DM currents may also produce radiated fields, but their impact is typically smaller due to the limited loop area associated with DM current paths. When the radiated fields originating from either CM or DM mechanisms couple into the antenna's reactive near field region, the local electromagnetic environment becomes distorted. This distortion modifies the effective gain pattern and can cause TRP reduction at particular angles.

In addition to pattern distortion, radiated noise is captured by the antenna as unwanted energy and is added to the desired receive signal. Any increase in the received noise level elevates the minimum input power required to meet the target BLER threshold, leading to degraded TIS performance. Radiated EMI therefore impacts both TRP and TIS through near field perturbation and direct noise coupling into the antenna.

6.9 Conducted Noise (150 kHz–30 MHz)

Conducted EMI in the 150kHz–30MHz band is mainly associated with DM currents generated by switching converters, the Power Distribution Network (PDN), and signal return paths. Fast switching transitions create voltage disturbances on the PDN that can couple toward the antenna port through parasitic impedances.

Although the dominant mechanism of conducted EMI is DM, CM components can emerge due to parasitic capacitances between switching nodes, the chassis, cable structures, and nearby grounding planes. These CM components at low frequencies may later convert into REs at higher frequencies.

When conducted noise reaches the antenna port, it directly increases the noise floor witnessed by the receiver. As receiver sensitivity is defined by the minimum received power needed to meet a target BLER, any elevation in conducted noise degrades TIS. Variations in PDN impedance can also slightly alter the reference impedance presented to the antenna, causing secondary deviations in TRP. Due to the high power converters and extensive PDN structures in EV chargers, conducted EMI is a key mechanism that affects both the receiver chain and the antenna interface.

6.9.1 Common Impedance Coupling as the Physical Origin of Conducted Noise

Conducted noise propagating through the PDN is fundamentally governed by the mechanism known as *common impedance coupling*. As shown in Figure 6.4, multiple circuits supplied from the same power source inevitably share parasitic impedances along

both the supply and return paths.

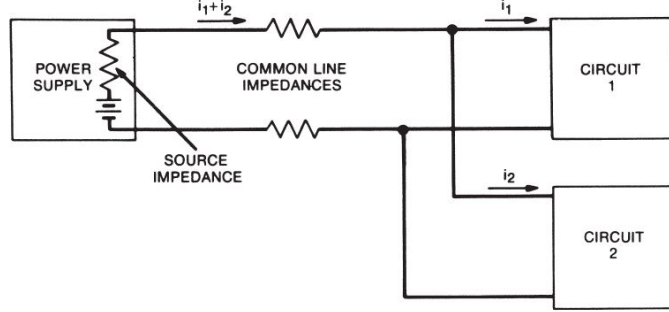


Figure 6.4: Common impedance coupling model illustrating the shared supply and return paths through which circuit currents interact [23].

These shared impedances, denoted collectively as Z_c , cause the voltage at the supply node to vary according to the total current flowing through the conductor. If the individual circuit currents are expressed as i_1 and i_2 , the resulting total current through the common impedance is

$$I_{\text{total}} = i_1 + i_2. \quad (6.18)$$

The corresponding voltage fluctuation appearing across the shared impedance is

$$V_c = (i_1 + i_2) Z_c. \quad (6.19)$$

This fluctuation directly modulates the supply voltage presented to each circuit. Consequently, a time varying current component $\Delta i_2(t)$ drawn by Circuit 2 appears as a conducted disturbance at the supply node of Circuit 1:

$$V_{\text{noise}, 1 \leftarrow 2} = \Delta i_2(t) Z_c. \quad (6.20)$$

This fundamental relationship, described by Ott [23], demonstrates that conducted EMI does not solely arise from switching transitions but is strongly shaped by the impedance

distribution of the PDN. Any variation in the shared current produces proportional voltage disturbances that propagate along the network and may ultimately reach sensitive RF nodes, including the antenna port. Therefore, common impedance coupling represents the primary physical mechanism through which low frequency conducted noise is injected into the receiver chain, forming the theoretical basis for understanding conducted EMI behaviour in complex electronic systems.

6.10 RF Desense and Its Impact on Antenna Performance

This section summarizes the combined effect of the electromagnetic noise mechanisms discussed in the previous subsections in the context of RF desensitization. RF desense refers to the degradation of the effective receiver sensitivity caused by unwanted electromagnetic interference coupling into the receiver chain.

The mechanisms examined in this chapter demonstrate that electromagnetic noise can reach the antenna and RF frontend circuits through both radiated and conducted paths. Radiated noise components may disturb the local electromagnetic environment within the reactive near field region of the antenna, leading to variations in antenna behavior. Conducted noise components, on the other hand, can be injected into the receiver chain through the PDN and common impedance coupling, resulting in an increased noise level observed at the receiver input.

RF desense is not the result of a single coupling mechanism, but rather a multifactor phenomenon arising from the combined influence of field coupling, current structure, frequency content, and system impedance distribution. Therefore, evaluating receiver performance degradation requires considering the interaction of multiple electromagnetic noise coupling paths rather than attributing the effect to a single dominant source.

The theoretical framework established in this chapter provides a conceptual basis for the experimental measurements and diagnostic analyzes presented in the subsequent

chapters, enabling a clearer interpretation of the physical origins of the observed performance deviations.



7. METHODOLOGY

This chapter presents the methodology used to diagnose and analyse the electromagnetic mechanisms that affect the cellular communication performance of AC EV chargers. The aim is to establish a clear, structured approach that links degraded TRP and TIS results to the physical sources of electromagnetic noise within the device.

The chapter first summarises the overall system architecture of the charger, and then introduces a step-by-step diagnostic framework that progressively narrows the analysis from system level behaviour to board level and component level interactions. This framework forms the basis for identifying dominant EMI mechanisms and evaluating their impact in the subsequent sections.

7.1 System Overview of AC EV Charger Architecture

AC EV chargers integrate power conversion stages, communication electronics, and mechanical structures within a unified embedded system. Before introducing the diagnostic methodology, a clear understanding of this architecture is essential. This is because each subsystem can influence the TRP and TIS performance through different physical mechanisms.

At a high level, an AC EV charger consists of three main subsystems: the power card (ACPW), the control card (ACCT), and the communication (base) card. The mechanical placement of these cards, the power distribution chain, and the communication links between them form the fundamental infrastructure for the electromagnetic evaluations presented in the subsequent sections.

7.1.1 *Power Card*

The ACPW board is the power input and power conversion card of the AC EV charger. This board is mounted on the back cover of the device and receives the AC mains

input. An internal power supply unit (PSU) located on the ACPW board converts the AC mains voltage into a regulated 12 V DC supply through an AC/DC conversion process. The generated 12 V DC voltage forms the main power rail of the system. This supply is first delivered to the ACCT board and is then distributed to the communication (base) card via the ACCT board using a flat cable. This architecture establishes a common power distribution chain for all control and communication subsystems within the device. The ACPW board is mechanically fixed to the back cover and serves as the central point for power generation and energy distribution within the overall system architecture.

7.1.2 Control Card

The ACCT board is the control card responsible for managing the charging operation scenarios of the AC EV charger. This board is mounted on the back cover together with the ACPW board and is mechanically positioned on bosses located on the ACPW board. The ACCT board is powered by the 12 V DC supply generated by the internal PSU on the ACPW board. Local step down converters on the ACCT board generate the required lower voltage levels for the MSP430 microcontroller unit (MCU) and associated control circuitry. The same 12 V supply is forwarded from the ACCT board to the base card via a flat cable, making the ACCT board an intermediate node in the system power distribution chain.

The ACCT board hosts an MSP430 MCU, which executes the main control logic responsible for charging initiation, charging operation, and charging termination. Communication between the ACCT board and the base card is established via a UART (Universal Asynchronous Receiver/Transmitter) interface, through which control commands related to the charging scenarios are transmitted.

7.1.3 Communication / Baseboard

The baseboard hosts the main processing and communication functions of the AC EV charger. This board is mounted on the front cover of the device. A chip antenna used for LTE communication is located on the base card and is positioned behind the front cover. Both the EVC01 and EVC08 platforms utilize the AM4378 system on chip (SoC) on the baseboard. In the EVC08 platform, this SoC additionally drives an Red Green Blue (RGB) display, whereas no display is present in the EVC01 platform. The base card integrates the LTE module, Wi-Fi module, Radio Frequency Identification (RFID) reader, and user interface peripherals. Communication with the LTE module is performed via a Universal Serial Bus (USB) interface, the Wi-Fi module is controlled over USB, the RFID reader is driven via an Serial Peripheral Interface (SPI) interface, and the device status LEDs are controlled using Pulse Width Modulation (PWM) signals. The baseboard receives its power from the ACCT board. The 12 V DC supply delivered from the ACCT board is converted into the required voltage levels for digital, RF, and peripheral components using on-board step down converters. Since both the LTE module and the antenna are located on this board, the base card constitutes the primary focus of the OTA performance evaluation.

7.2 Systematic Diagnostic Methodology for TRP and TIS Performance Issues

This chapter presents a systematic, measurement driven methodology for identifying the root causes of degraded TRP and TIS performance in AC EV chargers. The method relies on a sequential isolation of subsystems through conducted tests, antenna efficiency based estimations, baseboard OTA verification, and full device analysis. Each step answers one specific diagnostic question and produces measurable criteria that guide the subsequent stages.

7.2.1 Step 1: full device OTA Characterization

The diagnostic process begins with OTA testing of the complete AC EV charger. At this stage, the objective is to quantify the overall RF performance under realistic operating conditions and to determine whether the device meets the expected TRP and TIS requirements. The results indicate which LTE/5G bands deviate from the specification and whether the degradation is limited to specific angular directions or observed globally across all orientations. In addition, this step identifies whether the uplink path (TRP) or the downlink path (TIS) is the dominant limitation. Although this analysis provides a clear view of the performance degradation from a system perspective, it does not yet indicate the root cause; instead, it defines the frequency range, angular extent, and severity of the problem, and establishes the baseline for all diagnostic steps that follow.

The full product OTA evaluation is performed in an anechoic chamber using a 3D measurement system, as shown in Figure 7.1.

7.2.2 Step 2: Conducted Sensitivity and Conducted Output Power Measurements (Baseboard Level)

In this step, the LTE module is evaluated on the baseboard only, without the charger enclosure. The antenna remains inactive, and the module is connected to a Rohde Schwarz CMW500 through a coaxial cable inside a shield box. This setup eliminates radiated coupling and isolates the baseboard's intrinsic conducted performance, as shown in Figure 7.2.

Conducted sensitivity and conducted output-power measurements are performed to verify whether the baseboard layout, power delivery, or component placement introduce any degradation. If deviations are already observed at this stage, the source of degradation can be attributed to the baseboard itself, because no other subsystems are involved. In such cases, proceeding with full device OTA investigation would not be



Figure 7.1: An anechoic chamber is a structure whose interior surfaces are covered with pyramidal RF absorbers that attenuate electromagnetic waves and block external EMI. This allows antenna radiation patterns, TRP, and TIS measurements to be performed without multipath effects.

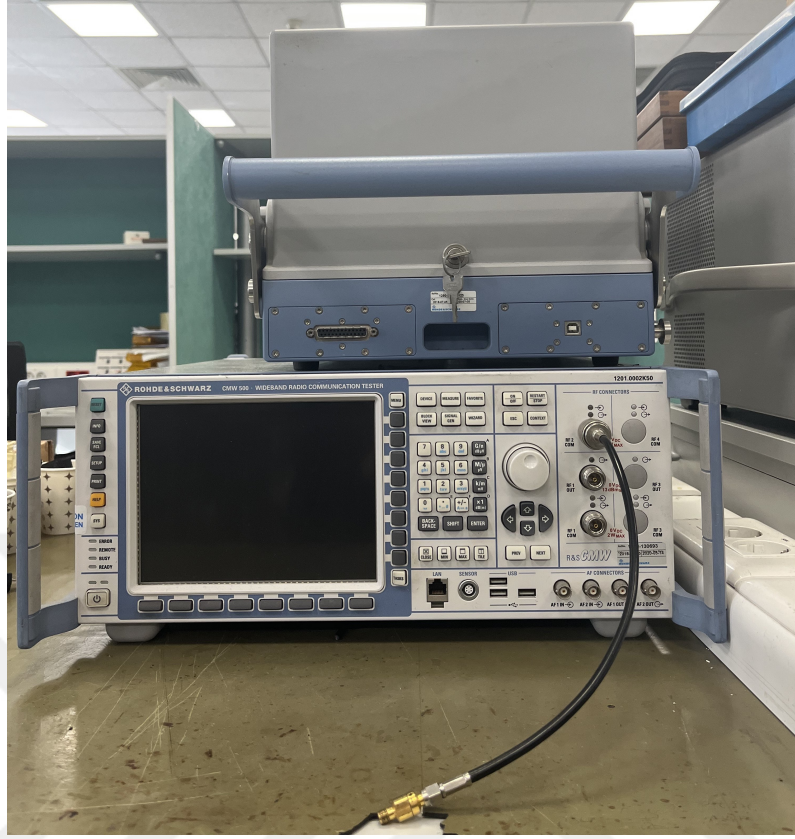


Figure 7.2: Conducted test configuration using CMW500 and shield box device.

meaningful until the baseboard issue is corrected. Conversely, if the conducted results are within specification, the RF modem and its interface are verified to be operating correctly, and the diagnostic process can move forward to radiated testing.

7.2.3 Step 3: Estimation of Expected TRP and TIS Using Conducted Measurements and Antenna Efficiency

To meaningfully evaluate the OTA behavior of the baseboard, the expected TRP and TIS values must be computed from conducted measurements and antenna efficiency η .

Antenna efficiency directly affects the expected TIS value. Since TIS is a degradative metric, lower efficiency increases TIS according to:

$$\text{TIS}_{\text{expected}} = P_{\text{sens,conducted}} - 10 \log_{10}(\eta) \quad (7.1)$$

Antenna efficiency also determines the expected TRP value. Since TRP is proportional to radiated power, lower efficiency decreases TRP:

$$\text{TRP}_{\text{expected}} = P_{\text{TX,conducted}} + 10 \log_{10}(\eta) \quad (7.2)$$

Equations (7.1) and (7.2) establish the ideal OTA performance that the baseboard should achieve in the absence of system level interference.

7.2.4 Step 4: Baseboard OTA Validation and Comparison with Expected Values

The LTE antenna is enabled again, and the baseboard is tested in an anechoic chamber. The measured OTA results are then compared with the expected TRP and TIS values obtained in the previous step 3. If the measured values are consistent with the expected ones, it indicates that the antenna, the RF path, and the baseboard design operate as expected from an electromagnetic point of view, and the root cause of performance degradation originates from the complete device, such as mechanical structures, switching circuits, or internal noise sources. If the measured values are significantly lower than expected, the problem is related to the baseboard itself, for example due to unintended loading, limitations in the RF layout, grounding problems, or effects related to power delivery. This step assumes that if the baseboard exhibits a performance problem, the same problem will eventually appear at the complete device level. Therefore, baseboard OTA validation is considered essential before conducting any further system investigation.

7.2.5 Step 5: Noise Sources Identification and Mitigation

This step aims to distinguish between radiated and conducted electromagnetic interference by selectively suppressing one coupling mechanism while evaluating the other. Radiated noise is assessed by minimising conducted noise paths through the use of an external low-noise power supply, allowing OTA performance to be evaluated primarily

under radiated coupling conditions. In contrast, conductance noise is investigated by reducing radiated coupling effects, enabling noise propagation through power and signal paths to be examined independently. This approach provides a systematic separation of radiated and conducted interference mechanisms and supports the identification of the dominant noise source. In addition to the classical distinction between radiated and conducted electromagnetic interference, near field EMI interactions constitute a separate coupling regime in compact electronic systems. In such cases, interference is not transferred through far field radiation or well defined conduction paths, but through localised electric or magnetic fields generated by internal noise sources.

Depending on the dominant field component, near field EMI may manifest as capacitive or inductive coupling between a noise source and a nearby susceptible structure. Consequently, near field EMI analysis is explicitly included within Step 5 of the proposed diagnostic methodology, enabling the identification of coupling mechanisms that cannot be classified as purely radiated or conducted.

7.2.6 Step 6: Radiated EMI Analysis

This step aims to determine whether the observed degradation in TRP and TIS is related to electromagnetic interactions between the antenna and the physical structure of the device. The analysis focuses on the electromagnetic environment surrounding the antenna and examines whether nearby mechanical or electronic structures alter the antenna operating conditions.

In particular, the impact of passive conductive structures located in close proximity to the antenna is evaluated to assess potential changes in antenna impedance, resonance characteristics, or radiation efficiency. This step establishes whether the performance degradation is associated with antenna placement and device geometry rather than electrical noise propagation through defined signal or power paths.

7.2.7 Step 7: Conducted EMI Analysis

This step evaluates whether the observed degradation of TRP and TIS is caused by noise propagating through electrical interconnections within the system. The analysis investigates whether interference is transferred as current along defined power or signal paths rather than through electromagnetic field interactions.

The evaluation includes the stability of power delivery to the LTE module, shared ground and return paths, and unintended coupling through PCB traces or cables. In addition, noise generated by switching power circuits and digital blocks is examined to determine whether it propagates into the module through power or signal lines and limits receiver sensitivity.

7.2.8 Step 8: near field EMI Coupling Analysis

This step addresses interference mechanisms that cannot be sufficiently explained by conducted noise propagation or far field radiation effects. In such cases, electromagnetic interference may arise through near field coupling, where localized electric or magnetic fields interact with nearby susceptible structures.

The analysis focuses on identifying capacitive and inductive coupling paths between internal noise sources and the antenna or adjacent conductive structures. High dv/dt switching nodes, high di/dt current loops, geometric proximity, floating conductive elements, and insufficient high frequency grounding are examined as key factors that may enhance near field coupling.

This step enables the identification of dominant near field coupling mechanisms and provides a physical basis for targeted mitigation strategies.

7.2.9 Step 9: Design Modifications and Iterative Verification

After the root cause is identified, the purpose of this step is to implement targeted design modifications and verify their impact on OTA performance. These modifications may include layout adjustments, improved grounding, repositioning of components, or refinement of the antenna placement. OTA measurements are then repeated to evaluate whether the applied modifications effectively improve TRP and TIS performance.

If the expected improvement is not fully achieved, the methodology does not terminate at this stage. Instead, the diagnostic workflow is intentionally taken back to Step 5, where the dominant electromagnetic mechanism is reclassified in light of the newly obtained results. This backward transition is not a mistake or unnecessary repetition, but a controlled and essential part of the methodology. In this way, both radiated and conducted mechanisms can be revisited when necessary before converging to the final solution.

The overall operation of the proposed diagnostic framework is summarised in the decision flow diagram presented in Figure 7.3. The diagram illustrates the logical progression between Steps 5–9 and explicitly highlights the iterative nature of the methodology, where design modifications are verified through repeated OTA measurements and earlier diagnostic steps may be revisited when required, until the desired TRP and TIS performance is achieved.

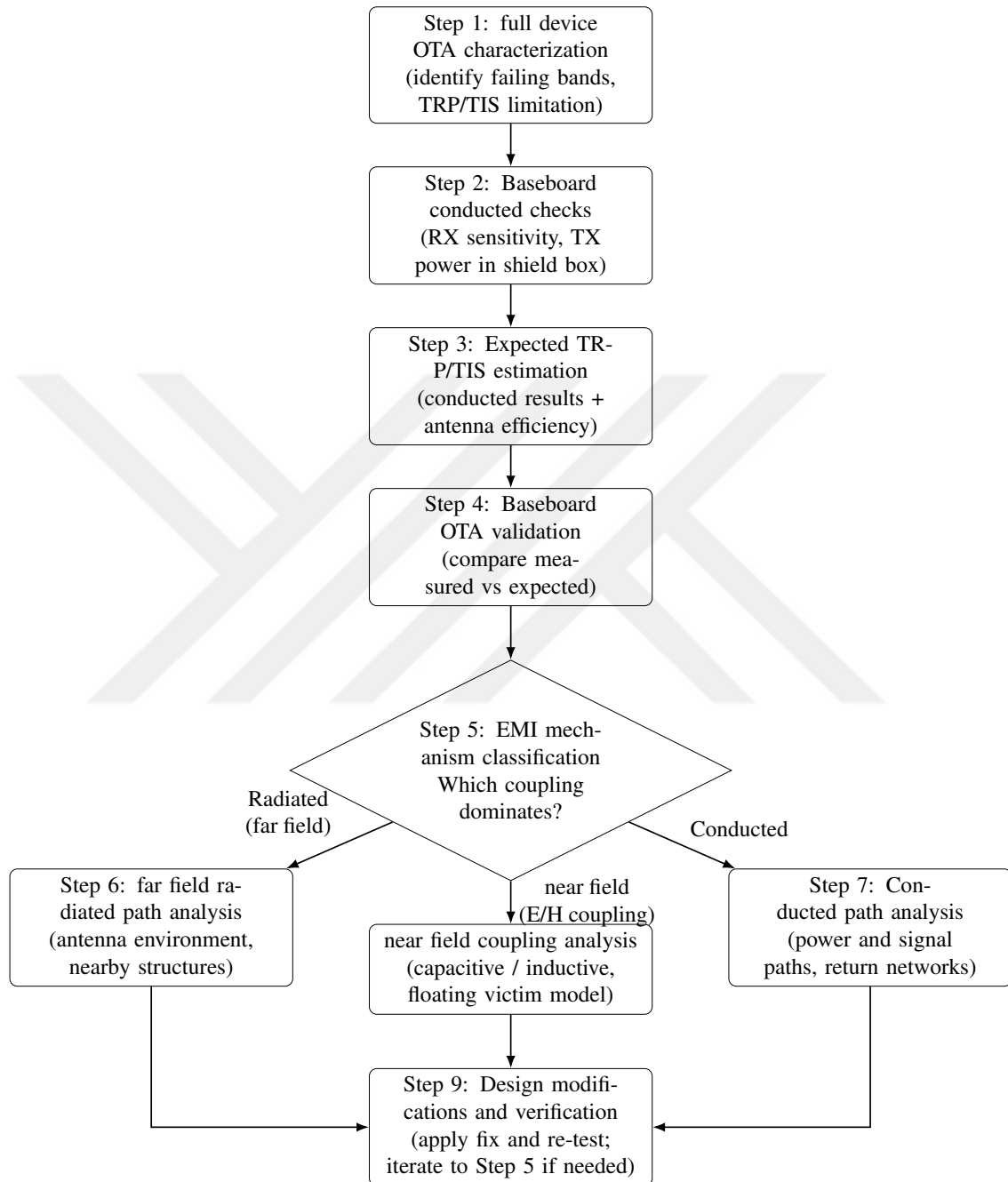


Figure 7.3: Systematic, measurement-driven diagnostic workflow for identifying the physical mechanisms responsible for TRP and TIS degradation in AC EV chargers.

7.3 Role of the EVC01 and EVC08 Platforms

The structured and iterative methodology proposed in this study was validated across AC EV chargers featuring different cellular communication architectures and usage scenarios. Within this scope, the EVC01 and EVC08 platforms were used as complementary test vehicles to demonstrate how the methodology operates under varying system level conditions. Both platforms are commercially available AC EV chargers that share a similar power electronics architecture, while exhibiting distinct differences in their communication subsystems and antenna configurations.

These differences enable the proposed methodology to be evaluated across a wide range of system complexities, from simplified single-antenna LTE architectures to multi antenna and Multiple Input Multiple Output (MIMO) based 5G systems, in a manner consistent with real world product development workflows.

7.3.1 EVC01 Platform

The EVC01 platform is designed for residential AC EV charging applications. It employs a Quectel EC25-EC LTE module, which provides both main and diversity antenna ports; however, only a single LTE antenna is actively used in the system configuration. This setup represents a simplified cellular communication architecture.

Owing to this simplified structure, the EVC01 platform allows the fundamental electromagnetic interactions affecting OTA performance to be observed and analyzed in a controlled manner. Measurements and analyses performed on EVC01 enable the identification of dominant electromagnetic mechanisms and potential design sensitivities that limit OTA performance. The findings obtained from the EVC01 platform serve as a reference for validating the core steps of the proposed methodology and provide critical technical input for subsequent product development stages.

7.3.2 EVC08 Platform

The EVC08 platform represents an AC EV charger intended for public charging applications. In its LTE configuration, the same Quectel EC25-EC LTE module is used; however, both the main and diversity antennas are actively employed. In addition, the system incorporates a Thin Film Transistor (TFT) Liquid Crystal Display (LCD) panel, a Measuring Instruments Directive (MID) unit, and a more complex *electronic integration*, which together increase the electromagnetic complexity of the antenna environment.

The EVC08 platform was designed taking into account the findings obtained during the OTA analysis of the EVC01 platform. In this way, key design sensitivities identified at the EVC01 stage were intentionally mitigated in the EVC08 design, allowing the effectiveness of the proposed methodology to be evaluated under a more complex and realistic communication architecture. This approach demonstrates that the methodology not only supports problem identification, but can also directly inform and guide product design refinement.

In the 5G configuration of the EVC08 platform, a fundamental change is introduced in the cellular communication architecture. The LTE module is replaced by a Quectel RM500U-EA 5G module, and a three antenna configuration is adopted in accordance with 5G MIMO principles. This configuration consists of two SMD monopole chip antennas and one external 5G FPC antenna. In this configuration, the panel, MID, and power architecture are kept identical to the LTE configuration; therefore, the electromagnetic environment is intentionally kept constant and only the impact of the change in the cellular communication architecture on the OTA performance is investigated.

Through this controlled architectural evolution, the proposed methodology is shown to be consistently applicable across different cellular technologies, ranging from single antenna LTE systems to multi antenna, MIMO based 5G systems. The evolution from the EVC01 platform to the LTE and 5G configurations of the EVC08 platform is summarized

in Figure 7.4.

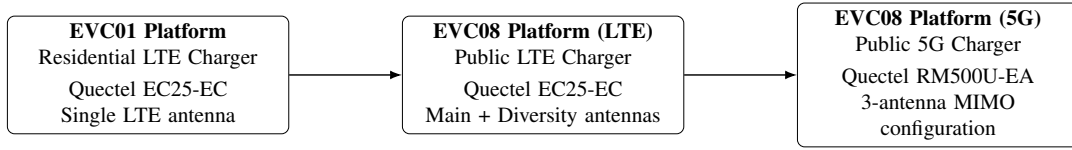


Figure 7.4: *Evolution of the experimental platforms used in this study, illustrating the transition from a simplified single antenna LTE architecture (EVC01) to a multi antenna LTE platform and finally to a MIMO based 5G communication architecture on the EVC08 platform.*

7.3.3 Applicability of the Methodology to 5G NR FR1 Systems

In contrast to the LTE-based evaluations, where the proposed methodology is used to identify and mitigate dominant electromagnetic interference mechanisms, the 5G NR FR1 systems evaluated in this study follow a different analysis focus. For the 5G NR FR1 configurations considered, the electromagnetic environment is treated as a known and previously characterized condition, and the analysis focuses on antenna configuration and MIMO related parameters rather than EMI root-cause identification.

Accordingly, a configuration-oriented evaluation flow is adopted for the 5G NR FR1 analysis, emphasizing passive antenna characterization, port–band mapping verification, MIMO path selection, and OTA validation prior to final configuration selection, as summarized in Figure 7.5.

The EVC08 platform provides a suitable test vehicle for demonstrating the applicability of this evaluation flow to practical 5G NR FR1 systems. The experimental results obtained using 5G OTA measurements are presented and discussed in Chapter 9.

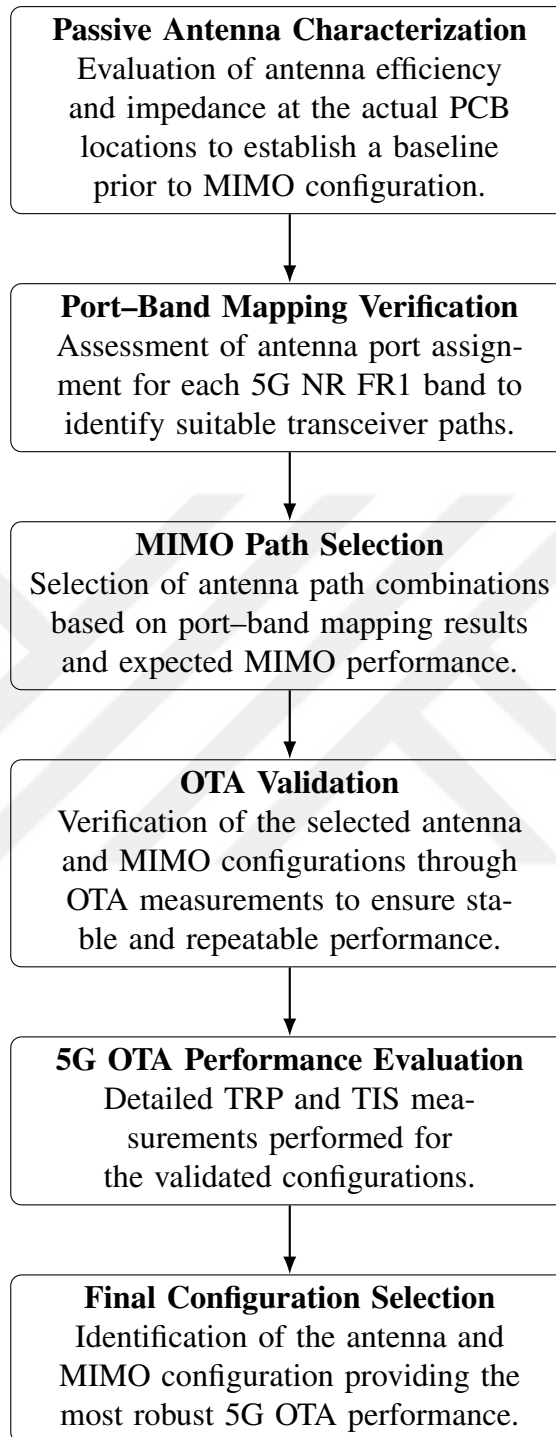


Figure 7.5: Configuration-oriented evaluation flow applied to the EVC08 platform for 5G NR FR1 analysis, illustrating the sequential steps from passive antenna characterization to final 5G OTA performance evaluation.

7.4 Common Hardware Architecture of the EVC Platforms

Although the EVC01 and EVC08 platforms differ significantly in terms of antenna number, communication module selection, and system integration, they are built on a shared hardware partitioning concept. In all three configurations considered (EVC01 LTE, EVC08 LTE, and EVC08 5G), the internal electronics are organized into three primary PCB level subsystems: the ACPW power conversion card, the ACCT control card, and a communication base board carrying the cellular module and antenna interfaces.

This common architectural backbone enables the proposed methodology to be applied systematically across different product generations. The EVC01 platform provides a simplified electromagnetic environment, allowing dominant coupling mechanisms to be clearly isolated and characterized. The EVC08 platform inherits the same control and power architecture, while evolving the communication subsystem first toward a dual antenna LTE implementation and then to a multi antenna 5G NR FR1 architecture. This progression therefore validates both the scalability and technology independence of the methodology.

Figure 7.6 summarizes this shared three-board partitioning and illustrates how the communication subsystem evolves across the evaluated platforms.

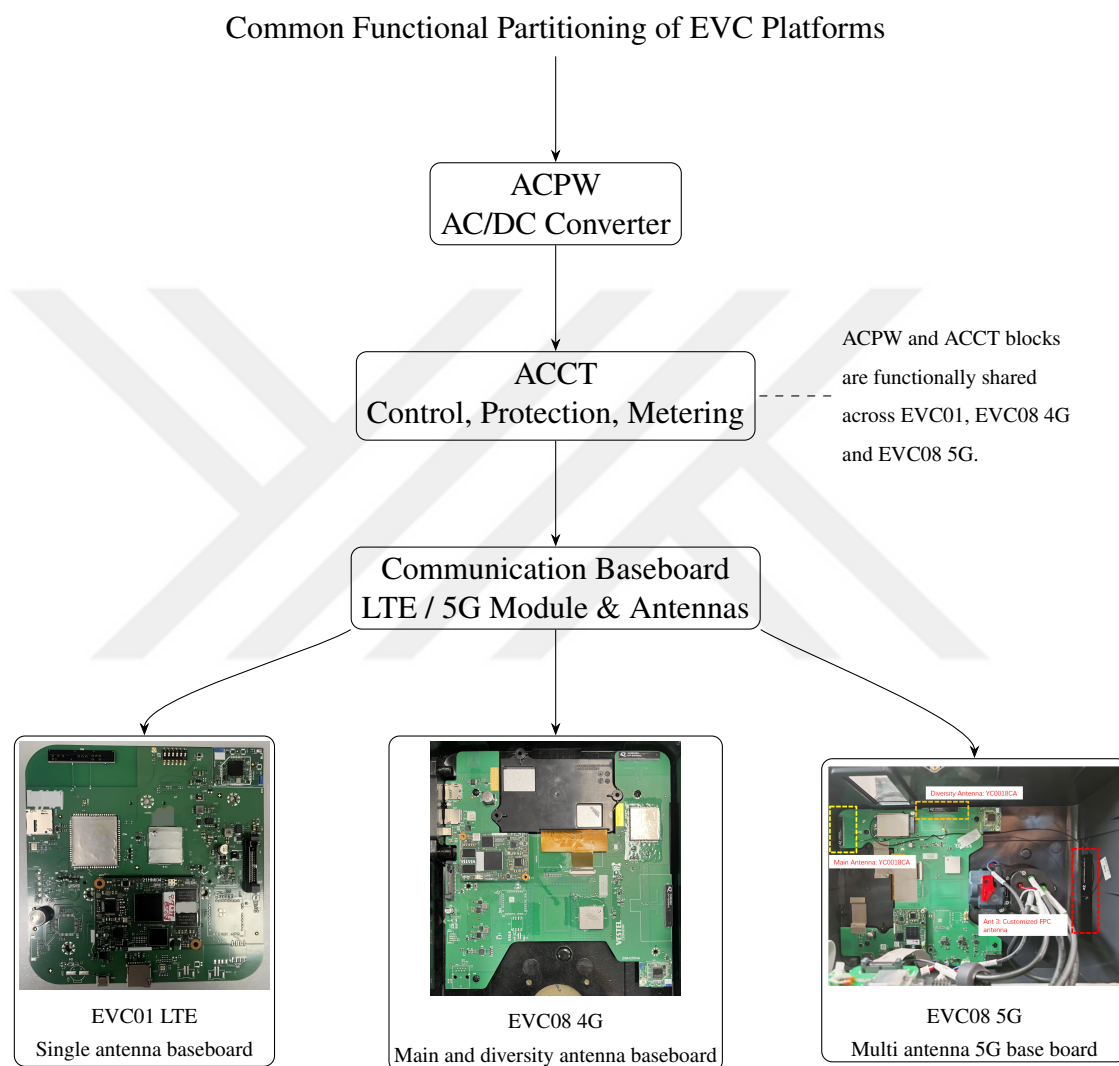


Figure 7.6: Common hardware partitioning of the EVC platforms and evolution of the communication baseboard from EVC01 LTE to EVC08 4G and EVC08 5G.

8. RESULTS

This chapter presents the OTA performance results of the AC EV charger platforms and evaluates them using the diagnostic methodology introduced in Chapter 7. All measurements were carried out in a fully anechoic chamber to ensure that the obtained TRP and TIS values reflect the device's intrinsic electromagnetic behavior rather than the properties of the test environment. In this way, the influence of internal structures, PCB layout, cable routing, and noise mechanisms on antenna performance can be reliably and controlled. For each platform, the results are examined in a sequence consistent with the proposed diagnostic methodology. These steps include evaluating the OTA performance at the device level, analyzing the isolation behaviour at the board level, identifying the dominant electromagnetic coupling mechanisms, and finally verifying the effectiveness of the applied design improvements. This structured evaluation helps distinguish whether the performance degradation originates from the antenna, the baseboard, or the device as a whole.

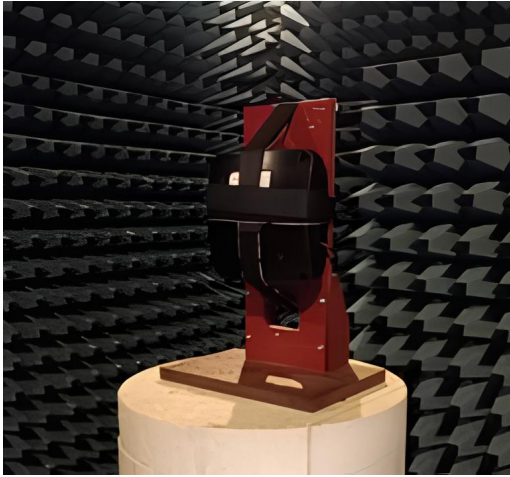
This chapter not only reports the TRP and TIS results, but also clarifies the physical and electromagnetic factors underlying the variations in performance. By relating the measured OTA behavior to the electromagnetic principles and interaction mechanisms discussed in the previous chapters, the obtained findings establish a consistent basis for the design recommendations and performance improvements presented in the following sections.

8.1 LTE OTA Performance Results for the EVC01 Platform

This section presents the LTE OTA performance results of the EVC01 platform and interprets them using the diagnostic framework defined in Chapter 7. The EVC01 platform is first analyzed as a reference case to demonstrate how the proposed methodology is applied in practice.

8.1.1 full device OTA characterization (Step 1)

The first stage of the experimental campaign corresponds to Step 1 of the diagnostic methodology and presents the OTA performance evaluation of the AC EV charger in full system operation. All internal subsystems, including the ACPW, ACCT, and base-board, are assembled in their mechanical configuration and powered through the internal power distribution architecture described in section 7.1. The EVC01 AC EV charger was placed inside the full anechoic chamber as the Device Under Test (DUT) and TIS values were measured for each channel (low, medium and high). The measured results are summarised in Figure 8.1. Figure 8.1(a) shows the EVC01 charger positioned inside the anechoic chamber during OTA testing, while Figure 8.1(b) presents the measured TIS results.



(a)

LTE	B20 dBm (TIS)	B8 dBm (TIS)	B28 dBm (TIS)
Low	-82.75	-81.49	-80.67
Mid	-82.88	-81.45	-80.39
High	-81.20	-81.77	-80.33
Average	-82.28	-81.57	-80.46
Spec	-86	-86	-86
Result	NOK	NOK	NOK

(b)

Figure 8.1: (a) EVC01 inside the full anechoic chamber during LTE TIS testing. (b) TIS results of EVC01 for LTE Bands B20, B8, and B28.

The measured TIS values exceed the limit -86 dBm in all all evaluated channels, indicating the out of spec sensitivity performance. Although all LTE bands supported by the module were measured, degradation is consistently observed in LTE Bands 8, 20 and 28. For this reason, the subsequent diagnostic steps focus on these bands to investigate

the underlying cause of the observed performance degradation.

8.1.2 Baseboard sensitivity under isolated conditions (Step 2)

Following the OTA measurements on the whole device that revealed the degradation of the TIS performance, Step 2 of the diagnostic methodology is performed. The purpose at this stage is to evaluate whether the baseboard itself contributes to any loss in receiver sensitivity. For this reason, the LTE module and the antenna are examined on the baseboard alone, without interaction from the remaining subsystems of the AC EV charger. Although the initial findings indicate that the entire device performs below the TIS requirement, they do not directly reveal whether the observed degradation originates from the baseboard itself or from system level interactions. Before reaching such a conclusion, a set of isolated measurements is required. As shown in Figure 8.2, the baseboard, which hosts both the LTE module and the antenna, is evaluated separately from the complete product to identify any potential issues on the board itself.

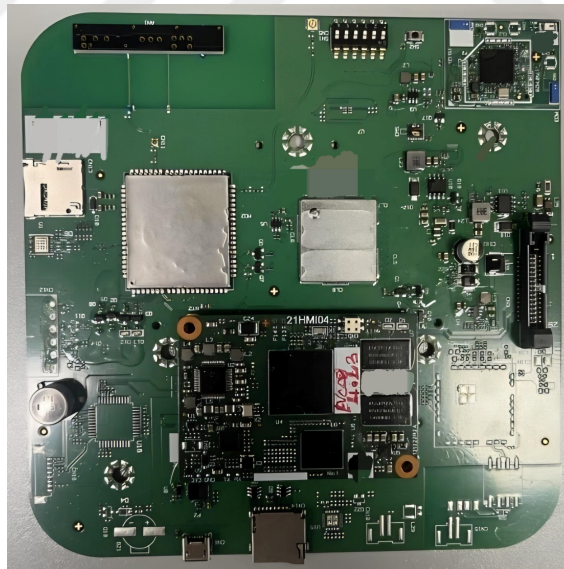


Figure 8.2: Baseboard containing the LTE module and antenna, tested independently from the full system.

The performance of TIS is primarily determined by two factors: the receiver sensitivity measured without the antenna (conducted sensitivity) and the efficiency of the antenna. The first step, therefore, is to perform conducted sensitivity measurements. The baseboard is tested with the antenna disabled, allowing the LTE module to be characterised independently of radiated effects. To avoid external disturbances, measurements are performed inside a shielded box, as shown in Figure 8.3.



Figure 8.3: Baseboard placed inside a shielded box for conducted receiver sensitivity measurements.

Table 8.1 summarises the sensitivity results conducted. For all LTE bands tested, the measured sensitivities meet the corresponding specification limits. This indicates that the LTE module, its power delivery network, and the associated routing on the baseboard meet the sensitivity requirements specified when the antenna is disabled. However, this configuration excludes antenna efficiency and radiated interactions, and therefore does not represent the OTA behaviour of the baseboard. The next step is to evaluate the baseboard inside the anechoic chamber with the antenna enabled and to compare the resulting OTA performance with the values expected from the conducted measurements.

Table 8.1: *Conducted sensitivity measurement results for LTE bands on the baseboard inside the shield box.*

LTE Band	Measured Avg. (dBm)	Spec (dBm)	Result
B20	-97.5	-97.5	OK
B8	-97.0	-98.5	OK
B28	-96.5	-98.7	OK

8.1.3 Expected TIS estimation and baseboard OTA validation (Step 3–4)

Since the conducted sensitivity results in Table 8.1 confirm that the LTE module operates within the specification limits, the next step is to evaluate the OTA performance of the baseboard with the antenna enabled. This corresponds to Step 3 of the diagnostic methodology, where the ideal TIS values are estimated based on conducted sensitivity measurements and antenna efficiency. According to (7.1), the expected TIS depends on the conducted receiver sensitivity and the antenna efficiency η . The LTE antenna used in this system is an SMD monopole chip antenna. Inspection of the manufacturer's datasheet shows that, in operating LTE bands, the average radiation efficiency is approximately $\eta \approx 0.25$. This value is therefore used in the following calculations. Using (7.1), the efficiency term is expressed as:

$$-10 \log_{10}(0.25) \approx 6 \text{ dB}.$$

Thus, the expected TIS for each LTE band can be obtained by adding approximately 6 dB to the conducted sensitivity value. Using the average sensitivities reported in Table 8.1, the expected resulting TIS values are:

$$\text{B20: } -97.5 + 6 = -91.5 \text{ dBm,}$$

$$\text{B8: } -97.0 + 6 = -91.0 \text{ dBm,}$$

$$\text{B28: } -96.5 + 6 = -90.5 \text{ dBm.}$$

These values represent the ideal TIS performance that the baseboard is expected to exhibit in the absence of system level interference. As in Step 4 of the methodology, the baseboard is tested with the antenna enabled, and the measured TIS values are compared with the theoretically calculated values. To this end, the baseboard is placed inside the anechoic chamber and TIS measurements are performed. Table 8.2 reports the measured values together with the expected TIS targets.

Table 8.2: *TIS performance of LTE bands for the baseboard in the anechoic chamber.*

LTE Band	Measured (dBm)	Expected (dBm)	Result
B20	-91.6	-91.5	OK
B8	-90.9	-91.0	OK
B28	-89.3	-90.5	OK

The close agreement between the expected and measured values confirms that the baseboard operates as intended when independently tested. Therefore, the reduced TIS observed during full device OTA testing must be associated with system level interactions, which are investigated in the next step.

8.1.4 Noise Sources Identification and Mitigation (Step 5)

This stage corresponds to Step 5 of the diagnostic methodology and aims to identify and classify sources of electromagnetic noise. Electromagnetic interference (EMI) can generally be categorised into two main types: radiated noise and conducted noise. To evaluate the effects of radiated noise in the AC EV charger, the device was tested in an

open cover configuration. During this test, the internal power conversion chain located on the ACPW board was bypassed, and the base board was supplied using an external regulated 12V DC adapter. This configuration excludes conducted noise mechanisms from the evaluation and enables a focused assessment of radiated noise emissions. The TIS results obtained under the open cover configuration (Table 8.3) are largely consistent with those measured when only the base board is installed (Table 8.2). These results indicate that radiated noise mechanisms are present within the AC electric vehicle charger system. To mitigate radiated noise, RF absorbing materials and electromagnetic shielding were applied around the antenna; however, these countermeasures did not result in a measurable performance improvement.

Table 8.3: *TIS performance of LTE bands for the AC EV charger with open cover in the anechoic chamber.*

LTE	B20 dBm (TIS)	B8 dBm (TIS)	B28 dBm (TIS)
Low	-90.50	-92.20	-89.90
Mid	-91.00	-91.90	-89.40
High	-90.72	-91.50	-89.30
Average	-90.74	-91.86	-89.20
Spec	-86	-86	-86
Result	OK	OK	OK

8.1.5 near field Investigation Under Radiated EMI Conditions(Step 6)

Following the identification of radiated electromagnetic interference (EMI) mechanisms in Step 5, the electromagnetic environment surrounding the antenna is examined in greater detail. When there is radiated coupling, structures located within the antenna's near field region may distort the electromagnetic field distribution and radiation pattern, thus degrading the antenna performance. For this reason, the next stage of the analysis focusses on the immediate mechanical environment of the antenna and the nearby internal structures defined by the device architecture.

According to Step 6 of the diagnostic methodology, the mechanical proximity between the antenna and the ACCT board is evaluated to determine whether the radiated

coupling inside the enclosure contributes to the observed TIS degradation. In the evaluated configuration, the ACCT board is positioned directly opposite the chip antenna, resulting in limited mechanical separation between the two structures. When the ACCT board is completely removed from the DUT, a measurable improvement in TIS is observed. However, no further improvement occurs when only the power to the ACCT board is disconnected and the board remains mechanically attached. Similarly, when only the bare ACCT PCB is present inside the enclosure, no improvement in TIS performance is observed. These observations indicate that the coupling mechanism is not primarily related to switching activity or conducted paths, but is instead associated with the physical presence of the PCB within the electromagnetic environment of the antenna.

The mechanical distance between the antenna and the ACCT board is measured as $d = 35$ mm. To assess the impact of this spacing, the relationship between the physical size of the antenna and the wavelength is considered. The antenna has a physical dimension of $D = 40$ mm.

For LTE Band 8 at 800 MHz, the wavelength is approximately $\lambda = 375$ mm. According to the electrically small antenna criterion given in (4.4), this condition is satisfied, since

$$40 \text{ mm} < \frac{375 \text{ mm}}{2\pi} \approx 59.7 \text{ mm}. \quad (8.1)$$

Therefore, the antenna is classified as electrically small at LTE Band 8. This result indicates that the reactive near field of the antenna extends approximately $\lambda/(2\pi)$, corresponding to about 59.7 mm. Since the physical separation $d = 35$ mm is smaller than this value, the ACCT board resides within the reactive near field region. In this region, even electrically small obstacles can perturb the field distribution and radiation pattern, leading to reduced antenna performance. This mechanism explains the increase in TIS values observed at low frequency LTE bands.

At higher frequencies, such as LTE Band 7 at 2.6 GHz, the situation changes. For this band, the wavelength is approximately $\lambda \approx 115.4$ mm, and the electrically small/large criterion in (4.5) yields the following result.

$$\frac{\lambda}{2\pi} \approx \frac{115.4 \text{ mm}}{2\pi} \approx 18.4 \text{ mm}. \quad (8.2)$$

Since the physical antenna size satisfies $D = 40 \text{ mm} > 18.4 \text{ mm}$, the antenna is classified as electrically large on the LTE Band 7. In this regime, the extent of the reactive near field can be estimated using (4.9).

Substituting $D = 40 \text{ mm}$ and $\lambda = 115.4 \text{ mm}$ into (4.9) yields

$$r_{\text{reactive}} \leq 0.62 \sqrt{\frac{(40 \text{ mm})^3}{115.4 \text{ mm}}} \approx 14.6 \text{ mm}. \quad (8.3)$$

The calculated reactive near field extent ($r_{\text{reactive}} \approx 14.6 \text{ mm}$) is significantly smaller than the physical separation $d = 35 \text{ mm}$. Therefore, at LTE Band 7 the antenna operates outside the reactive near field region with respect to the ACCT board. This geometric condition is consistent with the absence of a measurable TIS degradation in the high frequency LTE bands, in contrast to the behavior observed at lower frequencies. These findings confirm Step 6 of the diagnostic methodology, indicating that the degradation of TIS observed in low frequency LTE bands is primarily attributable to the reactive near field interaction between the antenna and the ACCT board.

8.1.6 Evaluation of Antenna Placement and Mechanical Integration

As concluded in Step 6, the antenna radiation pattern is affected by the presence of a PCB positioned directly in front of it, within its reactive near field region. Therefore, the analysis evaluates whether reducing the physical interaction between the antenna and the PCB can improve the TIS performance. To ensure that the improvements observed are solely associated with radiated coupling, all cutting experiments were conducted while

the base board was powered from an external 12V DC adapter. In this configuration, the internal Switching Mode Power Supply (SMPS) based AC–DC power conversion board was disabled, eliminating any switching noise from the power distribution path. Therefore, any improvement in TIS performance observed after cutting can be attributed only to the reduced electromagnetic interaction between the antenna and the PCB material. To experimentally verify this hypothesis, sections of the ACCT board facing the antenna are removed in different geometries. Two cut patterns are applied, as shown in Figure 8.4, corresponding to Case 1 and Case 2.

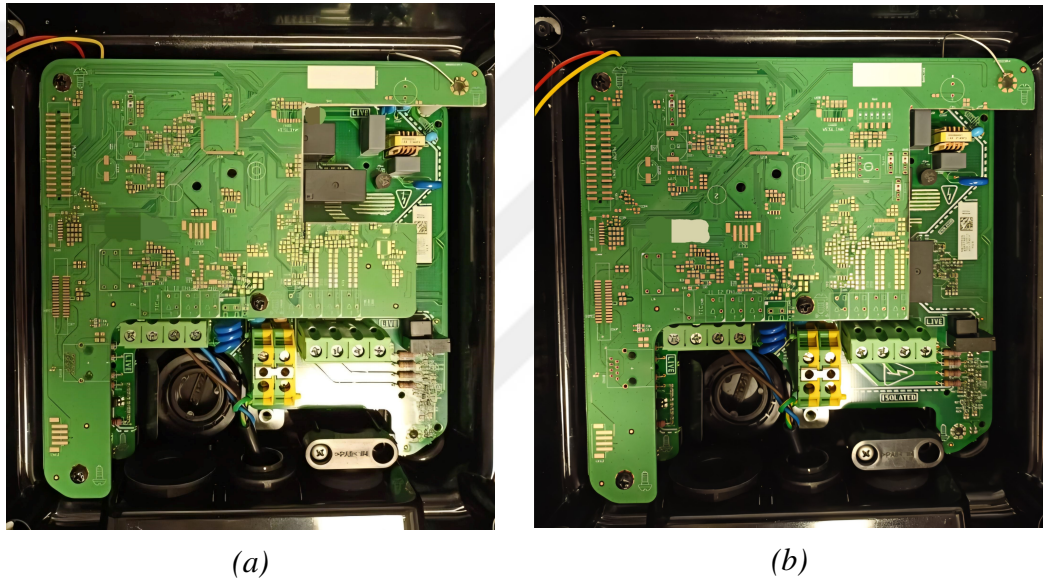


Figure 8.4: Cutting the ACCT board with different shapes to see the effect on TIS results
a) Case 1 b) Case 2.

The objective of these modifications is to reduce the effective surface area located within the reactive near field region, thereby minimizing reflections and parasitic coupling. To reduce the interaction between the antenna and the PCB, different cut geometries are applied to the ACCT board, as shown in Figure 8.4. The cut patterns are evaluated experimentally, and their corresponding TIS results are presented in Tables 8.4 and 8.5. Case 1 provides full compliance across all LTE bands, showing that removing the PCB material directly facing the antenna enables efficient radiation. In contrast,

Case 2 results in partial degradation only at Band 28, suggesting that the remaining PCB surface still introduces minor parasitic interaction. These findings confirm that reducing the physical area of the PCB within the reactive near field region leads to measurable improvements in TIS performance.

Table 8.4: *TIS results for modified PCB (Case 1 in Figure 8.4).*

LTE	B20	B8	B28A
Average (dBm)	-86.88	-88.82	-85.92
Result	OK	OK	OK

Table 8.5: *TIS results for modified PCB (Case 2 in Figure 8.4).*

LTE	B20	B8	B28A
Average (dBm)	-85.9	-87.69	-83.7
Result	OK	OK	NOK

8.1.7 Design Modifications and Iterative Verification (Step 9)

Based on the outcomes of the preceding investigation, a new ACCT board layout was developed in which the PCB regions directly facing the antenna were permanently removed to minimize interaction within the reactive near field region. The modified layout was generated using the Drawing Exchange Format (DXF) geometry obtained in the previous step and is shown in Figure 8.5. After fabrication, the modified ACCT board

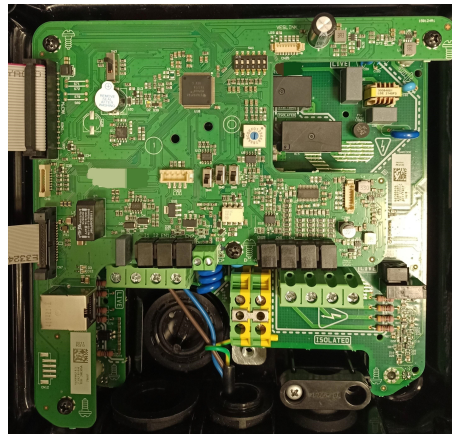


Figure 8.5: *Modified PCB layout designed to improve antenna performance.*

was reassembled into the complete AC EV charger and evaluated under nominal powering conditions to verify the OTA performance of the full system. The TIS results for the complete configuration are presented in Table 8.6.

Table 8.6: *TIS performance of LTE bands for the AC EV charger with the modified ACCT board in Figure 8.5.*

LTE	800 (Band 20)	900 (Band 8)	700 (Band 28A)
Channel	TIS (dBm)	TIS (dBm)	TIS (dBm)
Low	-82.35	-86.50	-82.33
Mid	-82.80	-87.30	-82.54
High	-81.76	-87.93	-82.65
Average	-82.30	-87.24	-82.51
Spec	-86	-86	-86
Result	NOK	OK	NOK

The results in Table 8.6 show that the TIS improvement observed under the PCB-only configuration is not fully maintained once the ACCT board is reinstalled and the charger is operated under its nominal powering conditions. Although the modified ACCT layout effectively reduces radiated interaction between the PCB and the antenna, this mitigation becomes insufficient when the full system is energized.

When the AC EV charger operates with its internal AC–DC power conversion path implemented on the ACPW board, switching activity associated with the internal power delivery network introduces additional electromagnetic disturbances. This effect is reflected in the approximately 3–4 dBm degradation observed in LTE Bands 20 and 28 when comparing the results in Table 8.6 with those obtained under the only PCB configuration.

These observations indicate that the performance degradation observed in the fully assembled system cannot be attributed solely to the PCB material positioned in front of the antenna. While the radiated coupling mechanism is effectively mitigated through the ACCT layout modification, additional noise mechanisms become dominant once the internal power delivery network is active. Therefore, the reduced performance in the full configuration cannot be attributed solely to the PCB located in front of the antenna. To

determine the relative impact of these mechanisms, the diagnostic workflow now returns to Step 5, where radiated and conducted paths are evaluated separately.

8.1.8 *Reclassification of Electromagnetic Mechanisms (Revisited Step 5)*

The results obtained in the previous steps showed that although reducing the PCB area in front of the antenna mitigated radiated coupling, TIS degradation reappeared when the system was powered through the internal AC–DC power conversion path on the ACPW board. This clearly indicated that a conducted interference mechanism was also present and could no longer be neglected. At this point, the diagnostic workflow was intentionally reverted to Step 5, not as a repetition, but as a necessary refinement of the earlier analysis, to identify the exact conduction path responsible for the observed degradation. In this stage, the presence of conducted noise was investigated using near field probes over the 0–30 MHz spectrum.

8.1.8.1 near field Identification of Conducted Noise Sources

In this step, near field probing was used to determine whether any switching activity inside the system couples into the RF front-end. The charger was operated in its nominal configuration and powered through the internal AC–DC power conversion path on the ACPW board to ensure that all system level switching activity was present during the measurements. The measurement setup is shown in Figure 8.6, where a magnetic field probe was scanned over different regions of the device while the spectral content was monitored on the oscilloscope. The resulting waveform captured from the probe is presented in Figure 8.7. The trace reveals several low frequency components within the 0–30 MHz range, which were consistently observed when the probe was positioned over the AC–DC power conversion circuitry on the ACPW board as well as over the ground region of the baseboard. These measurements show that the switching activity generated inside

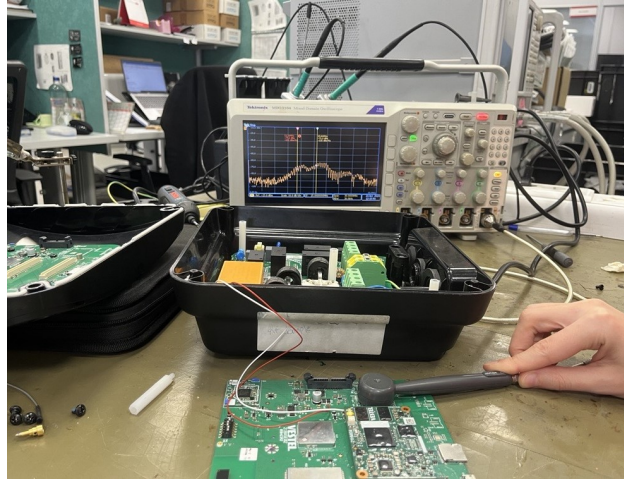


Figure 8.6: *near field probing setup used to examine switching activity inside the AC EV charger. A magnetic field probe is used to examine different regions while the device is powered through its internal PSU.*

the system is detectable at multiple locations, indicating the presence of a coupling mechanism. The exact nature of this coupling, whether radiated or conducted, is analysed in the next step.

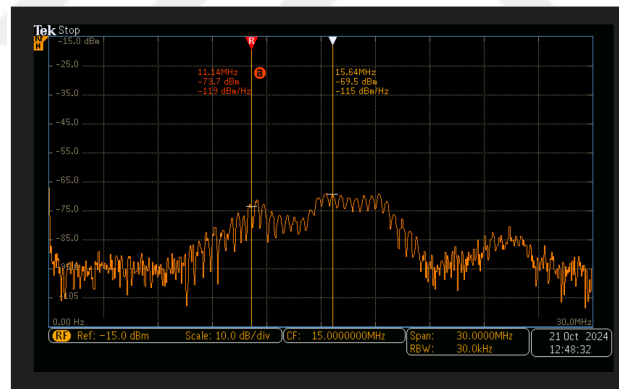


Figure 8.7: *Oscilloscope trace captured from the magnetic near field probe, showing low frequency switching noise components within the 0–30 MHz range. The probe is positioned over the ground pad region of the base board.*

8.1.8.2 Isolation Using External Supply and Radiated/Conducted Interpretation

To distinguish between conductive and radiative coupling mechanisms, the baseboard was disconnected from the internal power network and supplied from an external 12V adapter. During the test, the mechanical layout, antenna position and wiring remained unchanged; only the power delivery path was modified. Near field probing was repeated under the same conditions as in Step 7.1. Low frequency switching components were still observed when the probe was positioned over the AC–DC power conversion circuitry on the ACPW board. However, when the base board was powered from an external source, no corresponding spectral components were detected on the ground region of the base board. As shown in Figure 8.8, the ground pad remains noise free across the 0–30 MHz span. This confirms that the switching related disturbances reach the baseboard only when it is supplied through the internal power distribution path implemented on the ACPW board, and are no longer observable when the power distribution network is decoupled by external powering.

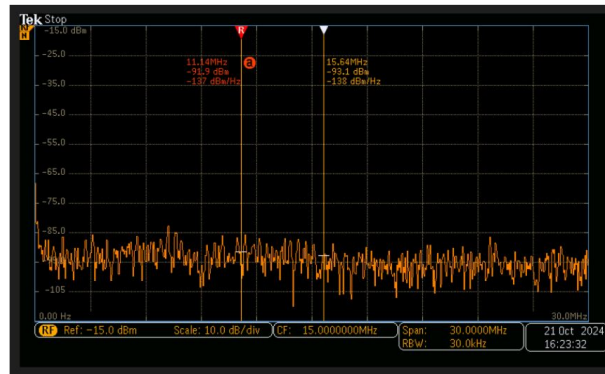


Figure 8.8: Oscilloscope measurement obtained from the ground pad of the baseboard using a near field probe when the board is powered from an external 12V supply. No spectral components appear in the 0–30 MHz range, indicating that the low frequency switching noise observed under internal PSU operation is suppressed when the common power distribution path is removed.

The absence of low frequency noise under external supply also rules out radiated

coupling. At these frequencies, the wavelength of the dominant switching component is

$$\lambda = \frac{c}{f} = \frac{3 \times 10^8 \text{ m/s}}{30 \times 10^6 \text{ Hz}} = \frac{3}{30} \times 10^2 \text{ m} = 10 \text{ m}. \quad (8.4)$$

The corresponding electrically-small threshold is

$$\frac{\lambda}{2\pi} = \frac{10}{2\pi} \text{ m} \approx \frac{10}{6.283} \text{ m} \approx 1.59 \text{ m}. \quad (8.5)$$

The largest physical dimension of the PCB is approximately

$$D \approx 150 \text{ mm} = 0.15 \text{ m}. \quad (8.6)$$

Substituting the numerical values into the electrically-small condition yields

$$D \ll \frac{\lambda}{2\pi} \Rightarrow 0.15 \text{ m} \ll 1.59 \text{ m}. \quad (8.7)$$

Equivalently, the normalized ratio becomes

$$\frac{D}{\lambda/(2\pi)} = \frac{0.15}{1.59} \approx 0.094 \ll 1, \quad (8.8)$$

confirming that the PCB is electrically small at $f \approx 30 \text{ MHz}$ and cannot act as an efficient radiator at these frequencies. Therefore, the observed low frequency switching noise cannot be attributed to radiated coupling and must instead be transported through conducted paths, propagating via the shared power and ground networks.

This observation directly motivates the interpretation of the measured behavior using the common-impedance coupling framework introduced in Section 6.9.1. In the investigated AC EV charger, both the ACCT and the baseboard are supplied through the internal DC power distribution network originating from the ACPW board. As a result, these subsystems share a common DC rail and ground return path characterized by a finite impedance Z_c .

As shown in Figure 8.9, the switching power stages implemented on the ACCT board draw a time varying current $i_1(t)$, while the communication (base) board draws its

own load current $i_2(t)$. Because both currents flow through the same shared power and ground impedance, the total current through the common supply path is given by

$$i_{\text{total}}(t) = i_1(t) + i_2(t). \quad (8.9)$$

The corresponding voltage fluctuation appearing across the shared impedance is

$$V_c(t) = i_{\text{total}}(t) Z_c \quad (8.10)$$

This fluctuation directly modulates the supply voltage and ground reference presented to both subsystems.

Consequently, a time varying current component generated by the ACCT card, denoted as $\Delta i_1(t)$, appears as a conducted disturbance at the supply node of the baseboard:

$$V_{\text{noise, base}}(t) = \Delta i_1(t) Z_c. \quad (8.11)$$

Figure 8.9 presents the interpretation of this mechanism at the block level for the investigated AC EV charger, while Table 8.7 summarizes how the classical common impedance coupling model corresponds to the charger architecture.

Table 8.7: *Correspondence between the classical common impedance coupling model and the AC EV charger architecture investigated in this study.*

Common Impedance Model	AC EV Charger Implementation
Power supply	ACPW based AC–DC power conversion
Shared impedance Z_c	Common DC rail and ground network
Circuit 1 (aggressor)	ACCT card (SMPS switching stage)
Circuit 2 (victim)	Baseboard
$i_1(t)$	Switching current of the ACPW card
$i_2(t)$	Load current of the base board
V_c	Conducted supply ripple at the baseboard

When the base board is supplied through the internal DC power distribution network shared with the ACCT board, the shared impedance path enables conducted noise propagation from the ACCT board to the base board through the common power and ground network. This coupling manifests as low frequency spectral components observed

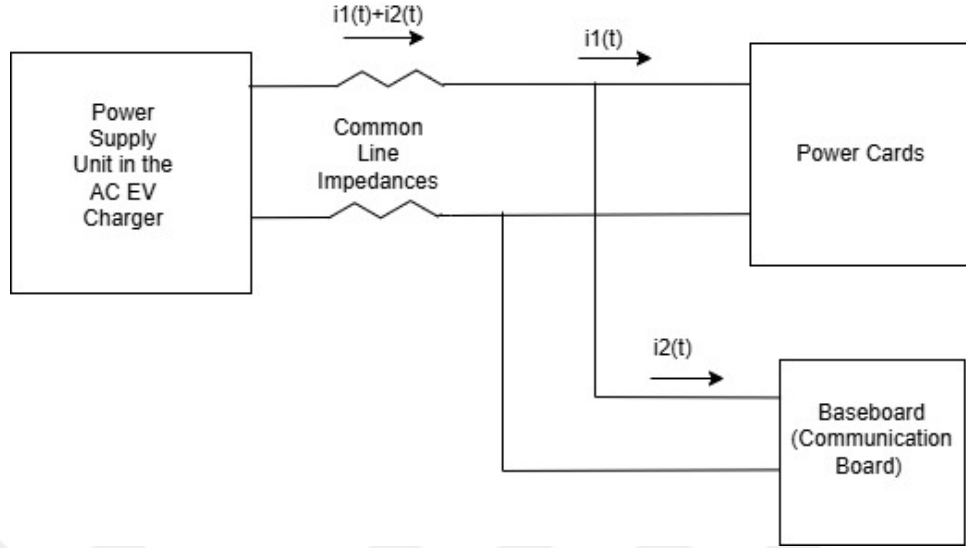


Figure 8.9: Block-level representation of common-impedance coupling in the AC EV charger. The internal PSU supplies both the ACPW card and the base card through a shared power and ground network with finite impedance Z_c , allowing switching-current-induced voltage fluctuations to appear on the baseboard.

on the ground region of the base board during near field probing. Conversely, when the base board is supplied from an external source, the shared impedance path is removed, and the corresponding low frequency spectral components are no longer observed on the base board. These results indicate that, under this configuration, the dominant coupling mechanism is governed by common-impedance coupling rather than by radiated effects.

8.1.9 Design Modifications and Iterative Verification(Step 9)

Since the expected improvement was not fully achieved in the first verification cycle, the workflow was intentionally taken back to Step 5, and the interference mechanism was reclassified based on the new OTA evidence. To reflect the combined impact of all design improvements introduced throughout the methodology, the analysis in this section corresponds to the final iteration stage described in Section 7.2.9. At this stage, the root cause of the OTA sensitivity degradation has already been identified through a combination of radiated interference analysis, near field probing, and analytical evaluation. The

focus of this section is therefore on verifying, via OTA measurements, how these targeted modifications perform when applied together under realistic system operation.

Earlier steps demonstrated that PCB cutting and layout modifications around the antenna region are effective in reducing radiated coupling mechanisms. However, OTA measurements performed under full system operation revealed that these improvements alone are not sufficient to fully restore sensitivity performance, indicating that additional coupling paths within the system must be considered. To isolate the influence of conducted noise during this final iteration, the test configuration was progressively restructured. First, the ACCT board, which is mechanically positioned in front of the antenna and previously identified as a radiated coupling source, was removed from the system. This configuration suppresses radiated interference paths and enables the evaluation of OTA performance primarily with respect to noise conveyed through the internal power distribution network.

Under this configuration, the base board was first powered using an external power supply, effectively disconnecting it from the shared internal power distribution path. For LTE Band 8, the measured TIS value was approximately -92 dBm. This result indicates that the antenna and RF receiver chain exhibit the expected reference OTA performance when both radiated and conducted effects are largely suppressed.

Subsequently, without modifying the mechanical or RF configuration, the base board was powered through the internal DC power distribution network originating from the ACPW board. Although the ACCT board remained removed, the measured TIS for LTE Band 8 degraded to approximately -84 dBm. This observation demonstrates that, even when radiated coupling paths are mitigated, noise conveyed through the internal power distribution network can adversely affect OTA sensitivity.

To mitigate this degradation, a CM filter was integrated into the base board power line while maintaining internal PSU operation. In this configuration, the measured TIS

for LTE Band 8 improved to approximately -87 dBm, restoring compliance with the TIS specification. The measured results for the three power delivery configurations are summarized in Table 8.8.

Table 8.8: *Measured TIS values for LTE Band 8 under different power delivery configurations during Step 9, with radiated coupling suppressed by removal of the ACCT board.*

Power Delivery Configuration	TIS (dBm)
Baseboard powered from an external DC source	-92
Baseboard powered through the internal DC power distribution network	-84
Internal DC power distribution with CM filtering applied	-87
TIS requirement (specification limit)	-86

Following this iterative verification, the ACCT board was reinstalled into the device in its modified form, as shown in Figure 8.5, where the section facing the antenna was removed to reduce radiated coupling. With the modified ACCT board integrated back into the system and the CM filter applied, OTA measurements were repeated under full system operation.

The final OTA results confirm that, for the LTE Bands 20, 8 and 28A, the TIS performance meets the specification requirements in all channels. The final TIS results obtained for the complete system configuration with CM filtering are presented in Table 8.9. These results demonstrate that desense behaviour in AC EV chargers cannot be

Table 8.9: *Final TIS performance of the AC EV charger for LTE Bands 20, 8, and 28A with CM filtering measured in the anechoic chamber under full system operation.*

LTE Channel	800 MHz (Band 20)	900 MHz (Band 8)	700 MHz (Band 28A)
Low	-84.65	-86.50	-85.35
Mid	-85.10	-87.30	-85.80
High	-84.06	-87.93	-84.76
Average	-84.60	-87.24	-85.30
Result	OK	OK	OK

attributed to a single coupling mechanism. Mitigating radiated coupling paths through mechanical and PCB-level modifications reveals the influence of noise conveyed through the internal power distribution network, while improvements applied solely to the power path are insufficient unless radiated effects are also addressed. This interaction confirms

that desense is inherently configuration-dependent and underscores the need for an iterative system level diagnostic methodology for reliable OTA performance improvement.



9. LTE OTA PERFORMANCE RESULTS FOR THE EVC08 PLATFORM

This section presents the LTE (4G) OTA performance results of the EVC08 platform by following the diagnostic methodology defined in Chapter 8. Compared to EVC01, the EVC08 AC EV charger features a more complex hardware configuration, including a diversity antenna, a TFT LCD panel, and a MID interface unit in addition to the main antenna, which significantly differentiates its electromagnetic environment from that of the EVC01 platform. Therefore, the measurements conducted on the EVC08 platform are of critical importance for evaluating the validity of the proposed methodology under different electromagnetic interaction mechanisms.

9.1 full device OTA Characterization(Step 1)

In this subsection, the LTE (4G) OTA performance results of the EVC08 AC EV charger obtained under full system operation are presented. Measurements were carried out in a CTIA-compliant full anechoic chamber with the device operating in its nominal configuration. The EVC08 test setup placed inside the anechoic chamber during OTA measurements is shown in Figure 9.1.

OTA measurements were conducted across all LTE bands supported by the EVC08 platform. However, degraded performance was observed in both TIS and TRP for LTE Band 8, and in TIS for LTE Bands 20 and 28. Therefore, the analysis in this section focuses on these bands. For each band, TIS and TRP measurements were performed at the standard low, medium and high channel frequencies defined in the OTA test plan. The baseline TIS and TRP results obtained for LTE Bands 8, 20, and 28 are summarized in Table 9.1. As shown in the table, both TIS and TRP values fall below the specification limits in LTE Band 8, while the TIS performance does not meet the acceptance criteria

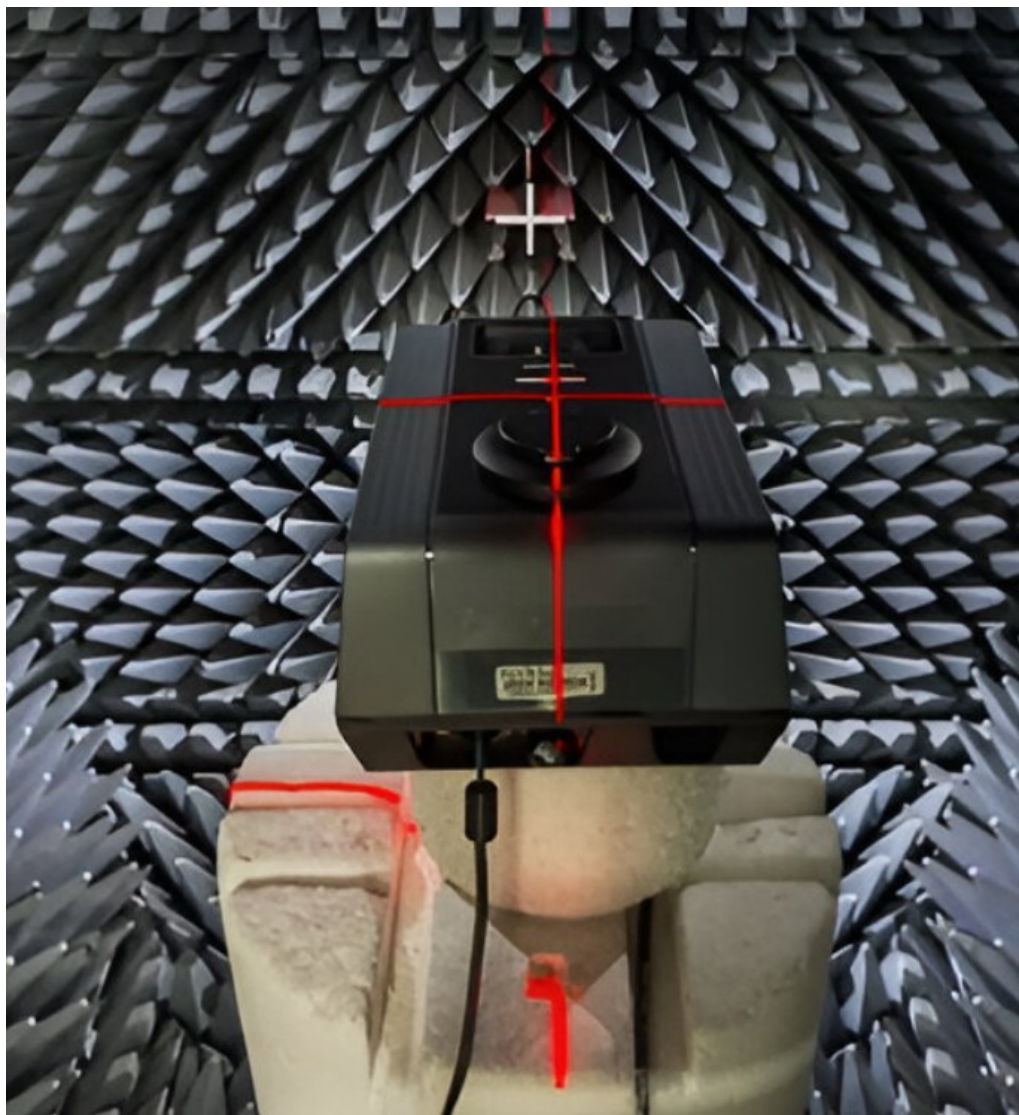


Figure 9.1: *EVC08 AC EV charger positioned inside the full anechoic chamber during LTE OTA measurements.*

in LTE Bands 20 and 28. Consequently, the OTA performance of the EVC08 platform is evaluated as *NOK* for all three bands.

Table 9.1: *Baseline TIS and TRP results of the EVC08 platform for LTE Bands 28, 20, and 8 measured in the anechoic chamber under full system operation.*

LTE Channel	B28 TIS (dBm)	B20 TIS (dBm)	B8 TIS (dBm)	B8 TRP (dBm)
Low	-80.27	-80.27	-83.96	16.01
Mid	-81.23	-81.23	-82.84	15.84
High	-81.04	-81.04	-84.75	15.90
Average	-80.85	-80.85	-83.85	15.92
Spec	-86	-86	-86	18
Result	NOK	NOK	NOK	NOK

These baseline results motivate the subsequent diagnostic steps, which aim to verify the behaviour of the baseboard and identify the dominant electromagnetic noise mechanisms responsible for the observed performance degradation.

9.2 Baseboard sensitivity under isolated conditions (Step 2)

Table 9.1 presents the baseline OTA results obtained for the EVC08 platform. The results indicate a pronounced degradation in both TIS and TRP, particularly in the low frequency LTE bands (Bands 8, 20 and 28). To investigate whether this degradation originates from system level interactions or from the communication hardware itself, the baseboard was isolated from the remaining subsystems of the AC EV charger and evaluated in a full anechoic chamber.

In the EVC08 platform, the baseboard constitutes an integrated structure that includes the LTE module, antennas, and the associated display interface. Therefore, the panel was retained as part of the baseboard configuration during this evaluation. The test configuration in which the baseboard was placed inside the anechoic chamber is shown in Figure 9.2.

The measurements performed under this isolated baseboard configuration exhibit

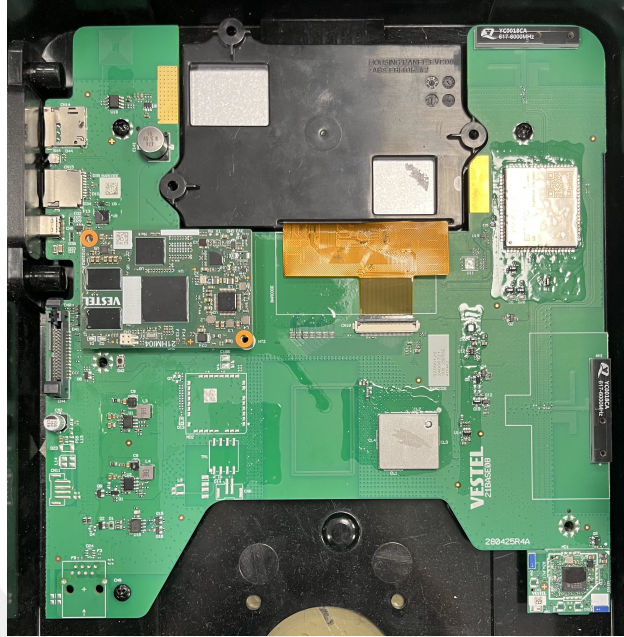


Figure 9.2: Baseboard of the EVC08 platform positioned inside the full anechoic chamber for isolated OTA evaluation.

a level of performance degradation comparable to that observed in the full device measurements. This observation suggests that the measured OTA performance degradation is not primarily caused by other system level components of the AC EV charger but may instead originate from the communication board environment where the antenna is integrated.

9.3 Noise Sources Identification and Mitigation(Step 5)

In this step, the diagnostic methodology is applied to distinguish whether the observed OTA performance degradation in the EVC08 platform originates from radiated or conducted electromagnetic interference (EMI) mechanisms. To this end, measurement results obtained under different test configurations are jointly evaluated in order to assess the dominant coupling path responsible for the observed degradation.

As an initial investigation, near field scanning was performed on the EVC08 baseboard to examine the potential influence of peripheral components on antenna performance. During the scan, a pronounced magnetic field activity was observed in the frequency range of 100–150 MHz, particularly on the metallic enclosure located at the rear side of the display panel. The oscilloscope output obtained from the near field probe measurement is shown in Figure 9.3.

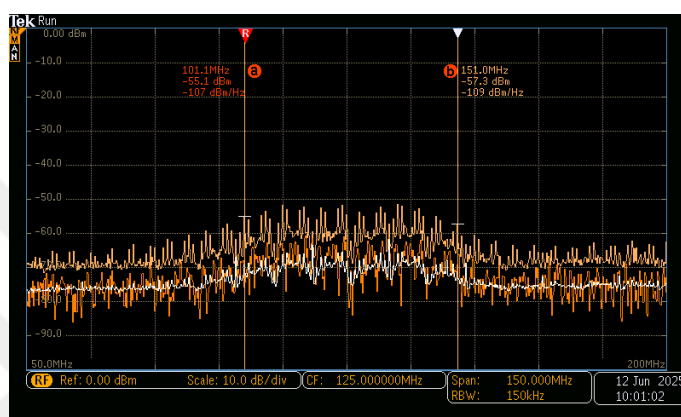


Figure 9.3: Oscilloscope output obtained from the near field magnetic probe measurement performed on the rear metallic enclosure of the TFT LCD panel, showing dominant activity in the 100–150 MHz frequency range.

Although this observation indicates that the panel is involved in the EMI process within the electromagnetic environment of the antenna, the near field measurement alone does not allow a definitive identification of the dominant interference mechanism. To verify this observation at the system level, the baseboard was subsequently tested in the anechoic chamber with the panel removed. The resulting OTA measurement results are summarised in Table 9.2. As shown in the table, a clear improvement is observed in the previously problematic LTE bands, and the measured performance satisfies the specified requirements. These results demonstrate a strong correlation between the presence of the panel and the observed degradation of OTA performance.

However, the fact that the OTA performance degradation persists even when the physical position of the panel is modified indicates that the observed interference cannot

Table 9.2: OTA performance results of the EVC08 platform measured in the anechoic chamber with the TFT LCD panel removed.

LTE Channel	B28 TIS (dBm)	B20 TIS (dBm)	B8 TIS (dBm)	B8 TRP (dBm)
Low	-86.0	-90.6	-92.4	18.18
Mid	-89.0	-89.2	-90.1	18.60
High	-90.2	-89.8	-90.3	18.69
Average	-88.4	-89.53	-90.93	18.49
Spec	-88.4	-86	-86	18
Result	OK	OK	OK	OK

be conclusively attributed to a purely radiative coupling mechanism. Nevertheless, the observed improvement alone is not sufficient to unambiguously classify the interference mechanism. For this reason, and in order to further investigate the nature of the observed interference, an EMC RE scan was performed on the EVC08 platform. The results of the EMC measurement reveal a distinct emission in the 100 to 150 MHz frequency range, consistent with the frequency content identified during the near field scanning. The result of the EMC RE measurement is presented in Figure 9.4. When considered together,

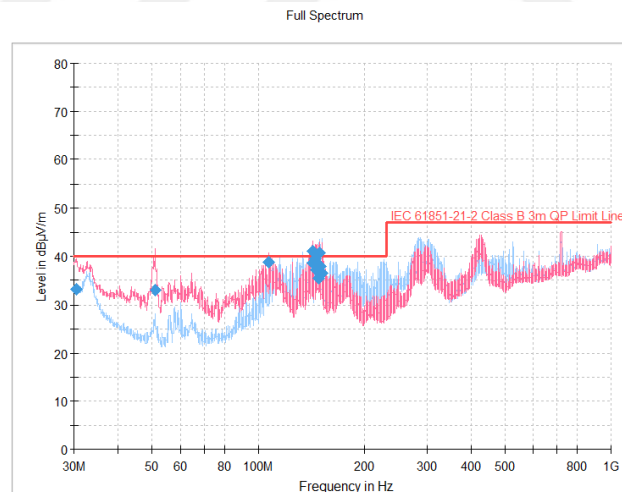


Figure 9.4: EMC RE measurement result of the EVC08 platform, indicating a distinct emission in the 100–150 MHz frequency range, consistent with the near field probe observations.

the OTA measurements, near field probe observations, and EMC RE results indicate the

presence of a common frequency component associated with the observed performance degradation.

9.3.1 Frequency-Domain Observation of the Problem

near field probe measurements and EMC RE scans performed within Step 5 reveal that the dominant electromagnetic activity in the system is concentrated in the **100–150 MHz** frequency range. When the TFT LCD panel is removed from the system, this frequency component disappears and a significant improvement in OTA performance is observed. This strong correlation indicates that the panel sub-system is directly involved in the observed performance degradation. At this stage, the primary objective is not to immediately apply a hardware modification, but rather to correctly identify the *physical coupling mechanism* through which the observed noise interacts with the antenna. Accordingly, the problem is evaluated within the classical electromagnetic interference framework, where noise mechanisms are typically classified as *conducted* or *radiated*.

9.3.2 Evaluation of Conducted EMI Possibility

In a classical conducted EMI scenario, noise is transferred as current through well defined conduction paths such as power lines, signal cables, or common impedance networks. In such cases, the interference mechanism is fundamentally associated with current flow along electrical interconnections. In the investigated configuration, the panel sub-system is powered and controlled via an FPC cable. However, experimental observations indicate that the dominant noise activity is not distributed along a conduction path, but instead appears to be localized on the metallic enclosure of the panel. This behaviour suggests that the observed interference is not governed by the transport of noise current along a defined conductor, but rather by a field related interaction at high frequencies.

Therefore, although electrical connections to the panel exist, the measured noise characteristics do not align with the expected behaviour of a classical conducted EMI

mechanism. As a result, conducted EMI is not considered the dominant coupling path in this case.

9.3.3 *Mathematical Assessment of far field Radiated EMI*

An alternative explanation for the observed interference is a far field radiated EMI mechanism. Under such a radiation scenario, the interfering structure is expected to behave like an antenna, with physical dimensions comparable to the operating wavelength.

To assess this possibility, the wavelength is calculated (4.1). For the dominant frequency range observed in Step 5 measurements (100–150 MHz), the corresponding wavelength lies approximately between 2 and 3 m. In contrast, the characteristic dimension of the panel metallic enclosure is approximately $L \approx 105$ mm. The resulting electrical size ratio is therefore

$$\frac{L}{\lambda} \approx 0.035 \text{ at } 100 \text{ MHz}, \quad \frac{L}{\lambda} \approx 0.0525 \text{ at } 150 \text{ MHz}. \quad (9.1)$$

These values indicate that the panel enclosure is electrically small within the frequency band of interest. Consequently, it is not expected to behave as an efficient far field radiator. This conclusion is consistent with the electrical size and field region analysis presented in Chapter 7. Accordingly, the observed interference cannot be adequately explained by a classical far field RE mechanism.

9.3.4 *Transition to near field EMI Interaction*

Since neither conducted EMI nor far field radiated EMI adequately explains the observed behaviour, the interference mechanism must be examined within the context of near field electromagnetic interactions. In this context, the term *near field* refers not to the antenna radiation regions, but to the electromagnetic fields generated by internal noise sources that interact with nearby conductive structures.

Digital signals present on the panel FPC, such as pixel clock and RGB lines, exhibit fast voltage transitions and therefore contain significant high frequency spectral components. These signals are capable of generating strong local electric fields, which may interact with nearby conductive structures within the device.

The metallic enclosure of the panel constitutes a large conductive surface located in close proximity to these noise sources. Although electrically connected to the system via the FPC, this connection does not provide a sufficiently low impedance path at the frequencies of interest. As a result, the panel enclosure is susceptible to near field electromagnetic interaction driven by local electric fields, motivating a more detailed coupling analysis in the subsequent section.

9.3.5 *Capacitive near field Coupling and Floating Victim Model*

The near field interaction identified in the previous subsection is dominated by electric field coupling, which can be described using a capacitive coupling model. Accordingly, the interaction between the digital noise source and the panel metallic enclosure is represented by an equivalent lumped-element network, as introduced in Chapter 7.

In this model, the noise source is coupled to the victim structure through a mutual capacitance C_{12} , while the victim structure is coupled to the system reference ground through a parasitic capacitance C_{2G} . The electrical connection between the victim structure and ground is represented by an equivalent resistance R , accounting for the finite impedance of the grounding path at high frequencies.

Under sinusoidal excitation, the induced noise voltage at the victim node V_2 relative to the source voltage V_1 can be expressed as

$$\frac{V_2}{V_1} = \frac{j\omega C_{12}}{\frac{1}{R} + j\omega(C_{12} + C_{2G})}. \quad (9.2)$$

When the victim structure is weakly grounded, the grounding impedance is large such

that

$$\frac{1}{R} \ll \omega(C_{12} + C_{2G}). \quad (9.3)$$

Under this condition, the resistive admittance becomes negligible compared to the capacitive admittance, and (9.2) reduces to

$$\frac{V_2}{V_1} \approx \frac{C_{12}}{C_{12} + C_{2G}}. \quad (9.4)$$

Equation (9.4) indicates that, in the high impedance (*floating*) case, the induced noise voltage at the victim node is governed primarily by capacitive voltage division. Consequently, even relatively small displacement currents can produce a significant noise voltage on the victim structure, as no low impedance path exists to effectively discharge the accumulated electric charge.

Based on the observations in Step 5, the panel metallic enclosure in the EVC08 platform exhibits this floating victim behaviour within the 100–150 MHz frequency range. The grounding path provided via the FPC does not constitute a sufficiently low impedance reference at these frequencies, making the capacitive near field coupling mechanism physically consistent with the measured EMI and OTA results.

9.4 Analytical Investigation of Capacitive near field Coupling (Step 8)

The findings obtained in Step 5 indicate that the observed interference in the 100–150 MHz frequency range cannot be consistently explained by purely conducted or far field radiated EMI mechanisms. Instead, the experimental observations point toward a near field coupling phenomenon. Accordingly, in this step, the interaction between the digital noise sources on the panel interface and the panel metallic enclosure is examined analytically using a capacitive near field coupling model.

The analysis is carried out in two stages. First, the spectral content of the identified digital noise sources is evaluated to establish their relevance to the observed frequency

band. Subsequently, the capacitive coupling model is analyzed under two limiting conditions corresponding to a weakly grounded (floating) victim structure and a well grounded (shielded) victim structure. This comparison allows the experimentally observed behavior to be associated with a specific analytical regime.

9.4.1 *Harmonic Content of the Digital Noise Source*

The panel is interfaced to the baseboard via a FPC that carries multiple digital signals, among which the pixel clock and RGB data lines exhibit fast voltage transitions and high switching activity. The nominal fundamental frequency of the pixel clock signal is approximately $f_0 = 9$ MHz.

For a periodic digital signal, the spectral components appear at integer multiples of the fundamental frequency, which can be expressed as

$$f_n = n f_0, \quad n = 1, 2, 3, \dots \quad (9.5)$$

The frequency range of interest identified in Step 5, namely 100–150 MHz, corresponds to harmonic orders

$$n = \frac{f}{f_0} \approx 11\text{--}17. \quad (9.6)$$

Thus, the measured emissions fall directly within the harmonic range of the pixel clock signal.

In addition to harmonic positioning, the magnitude of high frequency components is strongly influenced by the signal rise and fall times. A practical estimate of the upper frequency content of a digital signal is given by

$$f \approx \frac{0.5}{t_r}, \quad (9.7)$$

where t_r denotes the rise time. For rise times on the order of a few nanoseconds, which are typical for modern digital interfaces, spectral components extending into the 100–150 MHz range are physically expected. Therefore, the digital signals on the panel FPC

constitute a plausible noise source for the observed interference.

9.4.2 *Application of the Capacitive Coupling Model to the EVC08 Panel Structure*

Based on the above observations, the interaction between the digital noise source and the panel metallic enclosure is modeled using an equivalent capacitive near field coupling network. In this model, the coupling between the noise source and the metallic enclosure is represented by a mutual capacitance C_{12} , whereas the coupling between the metallic enclosure and the system reference ground is provided through a parasitic capacitance C_{2G} . The electrical connection of the metallic enclosure to ground is characterized by an equivalent high frequency impedance R , which reflects the effectiveness of the grounding path.

The analytical relationship describing the induced noise voltage on the victim structure relative to the source voltage has already been established in Equation 9.2. According to this expression, the coupled noise level is primarily governed by the values of C_{12} , C_{2G} , and the grounding impedance R . In the case of the EVC08 charger, the floating metallic panel corresponds to a configuration where C_{2G} is relatively small and the effective grounding impedance R is high, which enhances electric field coupling and leads to increased noise on the panel structure.

9.4.3 *Floating Victim Limit*

When the metallic enclosure is weakly grounded, its impedance to the reference ground is large, such that

$$\frac{1}{R} \ll \omega(C_{12} + C_{2G}). \quad (9.8)$$

Under this condition, the resistive admittance becomes negligible compared with the capacitive admittance, and the above expression reduces to

$$\frac{V_2}{V_1} \approx \frac{C_{12}}{C_{12} + C_{2G}}. \quad (9.9)$$

In this regime, the induced noise voltage is governed by capacitive voltage division and becomes effectively independent of the grounding impedance. As a result, relatively large voltage fluctuations may develop on the metallic enclosure due to displacement currents.

9.4.4 *Well Grounded Victim Limit*

In contrast, when the metallic enclosure is connected to the reference ground through a low impedance path, the following condition applies:

$$\frac{1}{R} \gg \omega(C_{12} + C_{2G}). \quad (9.10)$$

In this case, the capacitive term in the denominator becomes negligible, and the voltage ratio can be approximated as

$$\left| \frac{V_2}{V_1} \right| \approx \omega C_{12} R. \quad (9.11)$$

This expression indicates that the induced noise voltage on the victim structure is directly proportional to the grounding impedance. Consequently, reducing the effective high frequency grounding impedance leads to a proportional reduction in the coupled noise voltage.

9.4.5 *Numerical Illustration at 100 MHz*

To illustrate the difference between the two regimes, a numerical example is considered at $f = 100$ MHz. With $\omega = 2\pi f \approx 6.28 \times 10^8$ rad/s and representative parasitic values of $C_{12} = 5$ pF and $C_{2G} = 2$ pF, the floating case yields

$$\frac{V_2}{V_1} \approx \frac{5}{5 + 2} \approx 0.71. \quad (9.12)$$

Thus, in the floating regime, the induced noise voltage may reach approximately 71% of the source voltage.

In the well grounded regime, the same parameters give

$$\left| \frac{V_2}{V_1} \right| \approx \omega C_{12} R \approx 3.14 \times 10^{-3} R. \quad (9.13)$$

For $R = 10 \Omega$, this corresponds to approximately 31 mV, while for $R = 1 \Omega$ the induced voltage is reduced to approximately 3 mV. This comparison demonstrates that reducing the grounding impedance can suppress the induced noise voltage by more than an order of magnitude.

9.4.6 Analytical Implication for Design Mitigation

The harmonic analysis of the digital noise source, together with the capacitive near field coupling equations, indicates that the observed behavior in the EVC08 platform is consistent with the floating-victim regime. The comparison between the floating and well grounded limits further demonstrates that the effective high frequency impedance between the panel metallic enclosure and the reference ground is a dominant parameter in determining the magnitude of the coupled noise. From a physical perspective, capacitive near field coupling increases as the impedance of the victim structure with respect to ground rises. Therefore, reducing this impedance by establishing a well defined RF ground reference becomes a necessary mitigation strategy. In practical implementations, this is commonly achieved by connecting the conductive structure to the system ground at multiple points, thereby lowering its effective impedance across the frequency band of interest. This analytical implication directly motivates the design modifications introduced in the subsequent step.

9.5 Design Modifications and Iterative Verification (Step 9)

Step 8 analytical evaluations identified the primary interference mechanism as capacitive near field coupling between high frequency noise propagating along the FPC cable and a weakly referenced metallic panel structure. In this mechanism, the FPC cable acts as the primary noise source, while the metallic panel behaves as the affected structure (victim).

In the presence of capacitive near field coupling, the parasitic voltage induced on

the victim structure can be approximated by

$$V_{\text{panel}} \approx I_{\text{noise}} \cdot Z_{\text{panel}} \quad (9.14)$$

where I_{noise} represents the high frequency noise current flowing on the FPC cable, and Z_{panel} denotes the effective panel-to-ground impedance of the metallic panel.

In the initial configuration, the metallic panel was not directly bonded to the system ground and was only indirectly referenced through the FPC interface. Under this condition, the panel-to-ground impedance at high frequencies was dominated by parasitic reactive components rather than a well defined low-resistance return path:

$$Z_{\text{panel}}(f) = R_{\text{gnd}} + j\omega L + \frac{1}{j\omega C} \quad (9.15)$$

Due to the weak grounding reference, the effective grounding resistance R_{gnd} increased, leading to a significant voltage accumulation on the panel surface at MHz frequencies. This voltage accumulation resulted in a strong local electric field, thereby enhancing capacitive near field coupling.

Based on these findings, the first design modification introduced a single-point grounding connection between the metallic panel and the system ground. The purpose of this configuration was to establish a defined current return path, reduce the effective grounding resistance, and consequently lower the panel voltage level described in (9.14).

Following the implementation of single-point grounding, RE measurements were performed in the EMC chamber to verify whether the identified coupling mechanism was mitigated. The results indicated a reduction in emission levels at specific frequency ranges, confirming a partial suppression of voltage build-up on the panel structure. However, the single-point grounding configuration was insufficient to achieve a consistently low panel to ground impedance, particularly at lower frequencies.

Based on these observations, the metallic panel was connected to the ground of

the system at two spatially separated locations, implementing a dual-point grounding configuration. The practical implementation of this configuration is shown in Figure 9.5.

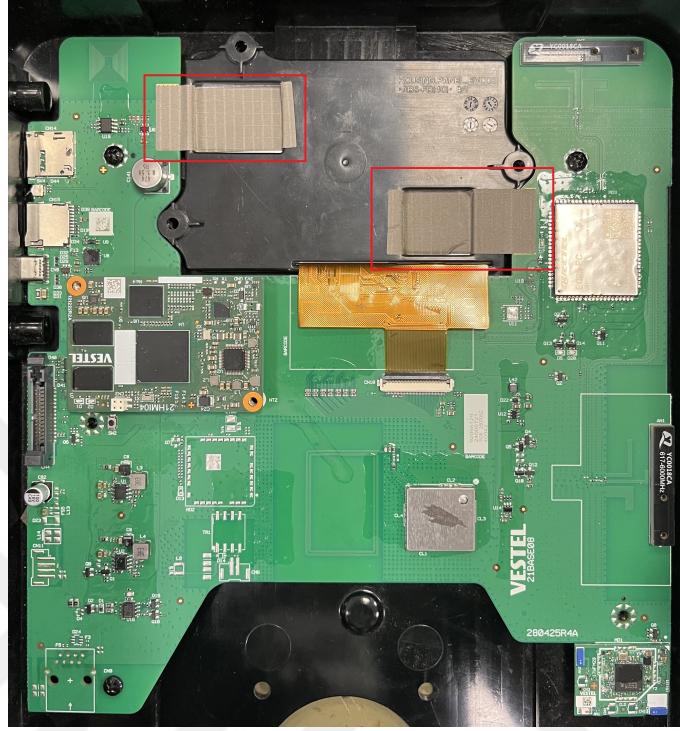


Figure 9.5: *Practical implementation of the dual-point grounding configuration, where the metallic panel is bonded to the system ground at two spatially separated locations to reduce the effective panel-to-ground impedance.*

By providing two parallel grounding paths, the resistive component of the panel-to-ground impedance is significantly reduced. As a result, the voltage buildup on the panel surface and the associated capacitive near field coupling are effectively suppressed:

$$Z_{\text{panel,dual}}(f) < Z_{\text{panel,single}}(f) \quad (9.16)$$

RE measurements were repeated after applying the dual-point grounding configuration. The EMC test results demonstrate a clear and consistent reduction in RE levels, as summarized in Figure 9.6. These results provide physical evidence that unintended current distributions and panel related near field effects were effectively suppressed.

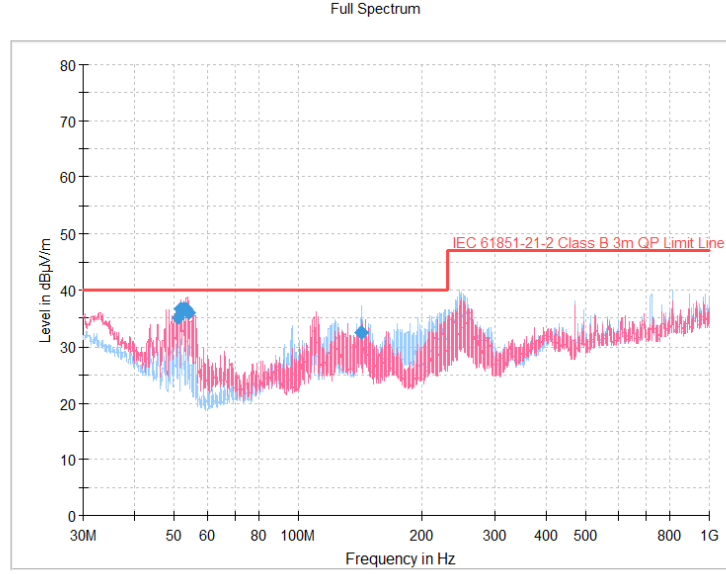


Figure 9.6: RE measurement results after applying the dual-point grounding configuration. A consistent reduction in emission levels is observed across the frequency range of interest, confirming the suppression of panel related near field effects.

9.5.1 OTA TRP Improvement and Resonance Suppression

Following the suppression of panel related near field effects verified by EMC measurements, OTA measurements were evaluated to assess the impact of the applied design modifications on antenna radiation performance. The results indicate a pronounced improvement in TRP, particularly for LTE Band 8.

LTE Band 8 operates in the uplink frequency range of 880–915 MHz. For the analysis, the center frequency was selected as $f_c = 897.5$ MHz. Accordingly, the free-space wavelength is calculated as

$$\lambda = \frac{c}{f_c} \approx 334 \text{ mm}. \quad (9.17)$$

The physical length of the metallic structure located behind the TFT LCD panel is approximately $L_{\text{panel}} \approx 105$ mm. This length exceeds $\lambda/4 \approx 85$ mm at the LTE Band 8 center frequency, indicating that the structure is electrically long enough to support resonance modes when its electrical boundary conditions are weakly defined.

In the initial configuration, the lack of a well defined high frequency reference to the system ground allowed standing-wave current distributions to form along the panel structure. These resonant currents increased the local electric field strength in the vicinity of the antenna, distorted the near field distribution, and degraded the measured TRP performance in the LTE Band 8. By providing two low impedance reference paths to the system ground, the applied design modification fundamentally altered the electrical boundary conditions of the panel. As a result, the minimum physical length required to support resonance shifted to approximately $\lambda/2 \approx 166$ mm. Since the panel length remained well below this threshold, panel related resonance modes were effectively suppressed, leading to the observed TRP improvement in LTE Band 8.

In addition to the modification related to the panel, the FPC cable connected to the display was mechanically fixed and referenced to the system ground, as shown in Figure 9.7.

This stabilisation reduced parasitic cable motion, unintended coupling, and noise-induced current fluctuations within the antenna near field region. As a consequence, receiver sensitivity improved, resulting in compliant TIS performance for LTE Bands 28, 20, and 8. As a final verification step, the complete device was placed inside the anechoic chamber and OTA measurements were repeated using the final hardware configuration. The measured TRP and TIS results for all evaluated LTE bands satisfied the specified requirements, as summarized in Table 9.3. These results confirm that the applied EMC and

Table 9.3: *full device OTA performance results of the EVC08 platform measured in the anechoic chamber using the final hardware configuration (Figure 9.7).*

LTE Channel	B28 TIS (dBm)	B20 TIS (dBm)	B8 TIS (dBm)	B8 TRP (dBm)
Low	-85.54	-85.54	-86.38	21.07
Mid	-85.58	-85.58	-85.49	20.96
High	-85.95	-85.95	-86.95	20.96
Average	-85.69	-85.69	-86.27	21.00
Spec	-86	-86	-86	18
Result	OK	OK	OK	OK

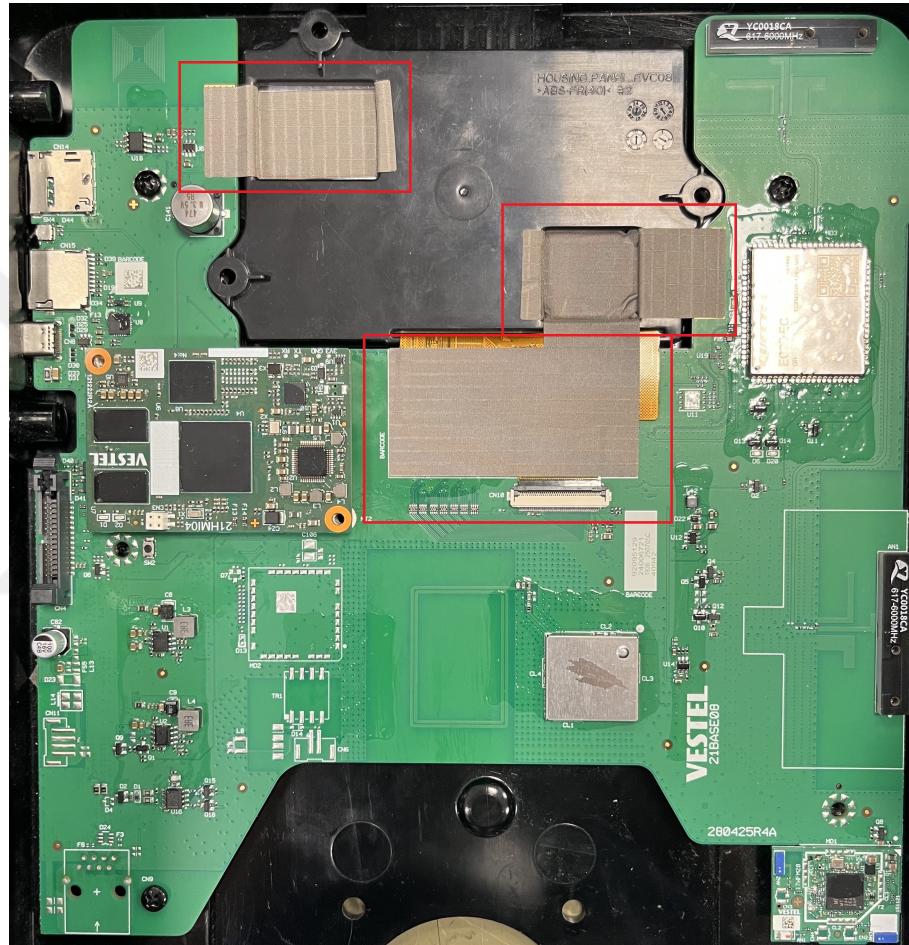


Figure 9.7: full device OTA performance results of the EVC08 platform measured inside the anechoic chamber with the final hardware configuration.

OTA-oriented design modifications collectively ensure compliant antenna performance at the full device level.

9.6 5G OTA Performance Results for the EVC08 Platform

This section summarizes the 5G OTA measurement setup and the test configurations evaluated on the EVC08 platform. All measurements were conducted following the configuration-oriented evaluation flow defined in Chapter 8. The 5G OTA measurements were performed on the EVC08 platform equipped with the Quectel RM500U-EA 5G NR FR1 module. The system supports a multi-antenna architecture consisting of two SMD monopole chip antennas and one external 5G FPC antenna. Throughout all measurements, the panel, MID unit, and power architecture were kept unchanged. The evaluated test configurations differ in terms of antenna port assignment, port–band mapping, and MIMO path selection. The subsequent sections present the results of passive antenna characterization, MIMO configuration analysis, and OTA performance measurements for the selected configurationse most stable and robust 5G OTA performance on the EVC08 platform.

9.6.1 *Passive Antenna Characterization Results*

Following the passive antenna characterisation, the next step focuses on verifying the port band mapping defined by the 5G NR FR1 module. In multi-port 5G transceivers, the association between frequency bands and RF ports is internally determined by the module architecture and cannot be arbitrarily modified at the system level. For the RM500U-EA module used in this study and as shown in the Figure 9.8, the port band mapping was obtained directly from the manufacturer datasheet and is summarised in Table 9.4.

For SMD chip antennas, manufacturers usually provide passive efficiency and



Figure 9.8: RF port configuration of the RM500U-EA 5G NR FR1 module used in the EVC08 platform.

impedance characteristics measured on standardised evaluation boards under ideal conditions. However, these reference measurements do not represent the actual electromagnetic environment of the final product, where the PCB layout, ground distribution, and nearby structures significantly affect antenna behaviour. The EVC08 platform used in this work features a PCB geometry and antenna placement that differ substantially from the reference evaluation boards provided by antenna manufacturers.

For this reason, all antennas were remeasured at their actual locations on the EVC08 communication board in a CTIA-compliant anechoic chamber. The final mechanical locations of the chip antennas and the external 5G FPC antenna within the product are shown in Figure 9.9.

The measured passive efficiency results clearly demonstrate that antenna placement and the surrounding PCB environment play a decisive role in determining passive

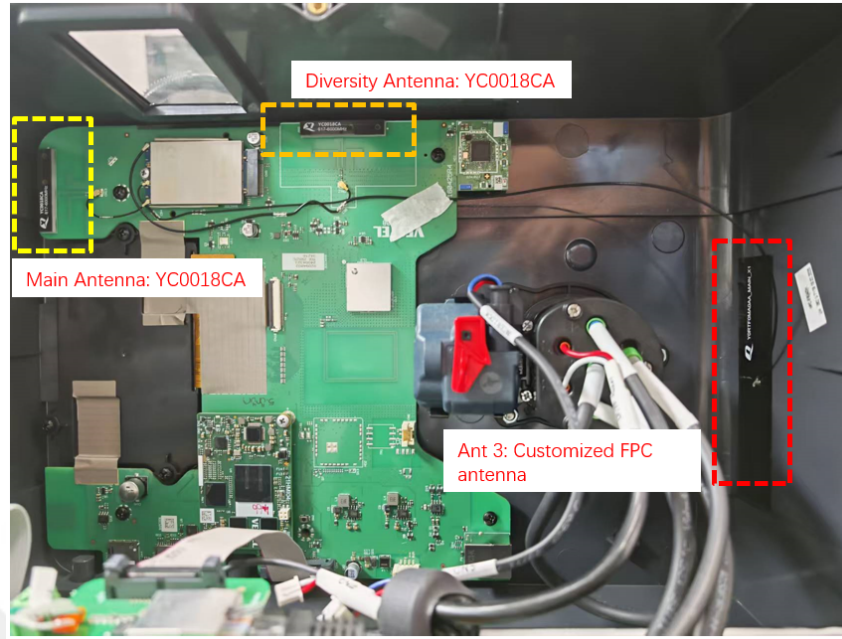


Figure 9.9: Final mechanical locations of the chip antennas and the external 5G FPC antenna inside the EVC08 enclosure.

antenna performance, as shown in Figure 9.10.

Based on these measurements, the antenna exhibiting higher passive efficiency was assigned as the main antenna, while the antenna with relatively lower efficiency was designated as the diversity antenna. For the external FPC antenna, multiple placement options within the enclosure were experimentally evaluated, and the location yielding the highest passive efficiency was selected for subsequent MIMO configuration analysis and OTA performance measurements.

9.6.2 Port-Band Mapping Verification

Following the passive antenna characterisation, the next step focuses on verifying the port-band mapping defined by the 5G NR FR1 module. In multi-port 5G transceivers, the association between frequency bands and RF ports is internally determined by the module architecture and cannot be arbitrarily modified at the system level. For the RM500U-EA module used in this study, the port band mapping was obtained

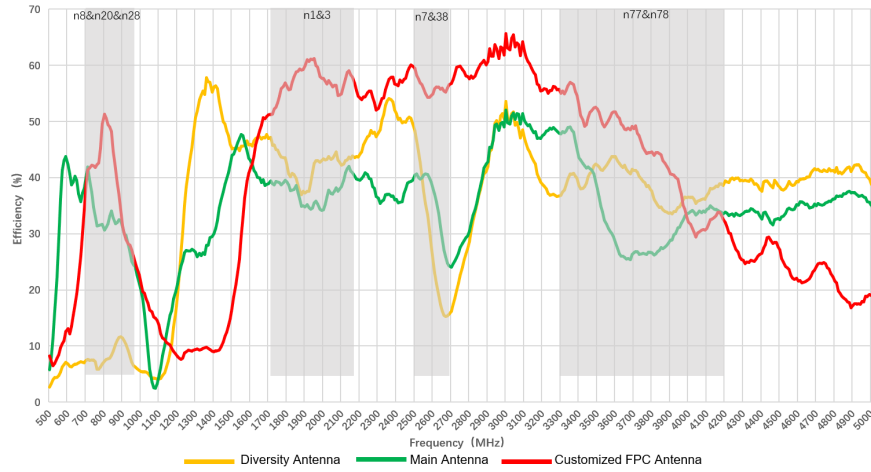


Figure 9.10: Measured passive efficiency of the antenna elements at their actual locations on the EVC08 base board, obtained in a CTIA-compliant anechoic chamber.

directly from the manufacturer datasheet and is summarized in Table 9.4.

Table 9.4: Antenna port–band mapping of the RM500U-EA cellular module extracted from the manufacturer datasheet.

Antenna	WCDMA / LTE	5G NR FR1 n1/n3/n7/n28/n66	5G NR FR1 n5/n8/n20	5G NR FR1 n38/n40/n41	5G NR FR1 n77/n78
ANT0	LMHB TRX	PRX MIMO	TRX	TRX1	TRX1
ANT1	–	DRX	–	DRX	DRX
ANT2	LMHB DRX	DRX MIMO	DRX	DRX MIMO	DRX MIMO
ANT3	–	TRX	–	TRX0	TRX0

The mapping therefore establishes the fixed antenna–port associations that form the basis for all subsequent OTA and MIMO evaluations. To support multi-antenna operation, the RM500U-EA module defines multiple receive paths. The primary receive path (PRX) represents the main downlink reception branch associated with the active transmit receive path (TRX), while the diversity receive path (DRX) provides an additional spatially separated receive branch to improve downlink robustness. For bands supporting MIMO operation, PRX MIMO and DRX MIMO indicate that multiple receive ports are simultaneously active within the MIMO receiver architecture.

As the TRX path directly defines the transmit power and the primary downlink receive branch, the antenna assigned to this path fundamentally determines both the TRP and the baseline TIS performance. Additional receive paths such as PRX and DRX can improve the robustness of TIS only when the TRX antenna provides a sufficiently strong baseline. If a low-performance antenna is assigned to the TRX path, neither diversity nor MIMO reception can compensate for the resulting degradation in OTA performance.

9.6.3 5G OTA Results — Antenna Reconfiguration Scenarios and EMI Interaction

Following the passive antenna characterisation, four different antenna configurations were evaluated to investigate the combined impact of EMI and port–band mapping on OTA performance. In each configuration, TRP and TIS were measured for the n5, n8, n20, and n28 bands. The results revealed that both antenna placement and port assignment significantly influenced system performance through their interactions with internal EMI sources.

9.6.3.1 Configuration 1 — TRX SMD Main Antenna (n5/n8/n20 Bands) and TRX External Main Antenna (n28 Band)

In the initial configuration, the SMD main antenna with higher passive efficiency was connected to the ANT0 port (TRX path), while the weaker antenna was connected to ANT3 (PRX path). As shown in Table 9.5, the TRX path corresponds to ANT0 for the n5, n8, and n20 bands, and to ANT3 for the n28 band.

Table 9.5: Measured TRP and TIS values for Configuration 1.

Band	TRX Path	PRX/DRX Paths	TRP (dBm)	TIS (dBm)
n5/n8/n20	ANT0 (SMD Main)	DRX: ANT2 (External)	20.2	−65.8
n28	ANT3 (External)	PRX: ANT0 (SMD Main), DRX: ANT2 (External)	20.3	−81.1

Within the charger enclosure, the MID (metering) unit was positioned directly opposite the TRX antenna, acting as a dominant EMI source during charging. For the n5,

n8, and n20 bands, TRP was approximately 20.2 dBm, whereas TIS decreased to around -65.8 dBm. This degradation was mainly attributed to EMI coupling between the MID circuit and the TRX antenna, as the downlink reception relied solely on the TRX path and no diversity support was available. In contrast, for the n28 band, the TRX path corresponded to the external antenna, which was electromagnetically decoupled from the EMI region. As a result, stable performance was achieved with TRP of 20.3 dBm and TIS of -81.1 dBm. These findings demonstrate that the degradation at the lower bands was primarily driven by TRX–EMI interaction rather than passive antenna efficiency.

9.6.3.2 Configuration 2 — TRX External Main Antenna (n5/n8/n20 Bands) and TRX SMD Main Antenna (n28 Band)

In the second configuration, the external antenna was assigned to the TRX path for the n5, n8, and n20 bands, while the SMD main antenna was assigned as the TRX path for the n28 band. The corresponding TRX and PRX/DRX assignments and the measured OTA results are summarised in Table 9.6.

Table 9.6: *Measured TRP and TIS values for Configuration 2.*

Band	TRX Path	PRX/DRX Paths	TRP (dBm)	TIS (dBm)
n5/n8/n20	ANT0 (External)	—	20.2	-76.9
n28	ANT3 (SMD Main)	PRX: ANT0 (External)	20.7	<i>Disconnected</i>

For the n5, n8, and n20 bands, assigning the external antenna to the TRX path resulted in a TRP of approximately 20.2 dBm, while the TIS improved to -76.9 dBm. Compared to Configuration 1, this result indicates a partial recovery of downlink performance once the TRX antenna was relocated further away from the dominant EMI region inside the charger enclosure. Since no diversity paths were active in these bands, the observed performance improvement can be directly attributed to the change in the TRX antenna position rather than any MIMO or diversity gain. In contrast, for the n28 band, the TRX path corresponded to the SMD main antenna, while the external antenna acted

as the PRX branch, as shown in Table 9.6. Although the TRP was measured as 20.7 dBm, the receive path was effectively disconnected during the measurement, preventing a valid TIS result from being obtained. This condition indicates that the port–band mapping and antenna assignment in Configuration 2 did not provide a stable receive architecture for the n28 band.

9.6.3.3 Configuration 3 — TRX External Main Antenna (n5/n8/n20) + DRX SMD Main, and TRX SMD Diversity Antenna (n28)

In the third configuration, the external main antenna remained on the TRX path for the n5, n8, and n20 bands, while the SMD main antenna was assigned to the DRX path to enable receive diversity. For the n28 band, the TRX path corresponded to the SMD diversity antenna, the PRX path to the external antenna, and the DRX path to the SMD main antenna. The corresponding TRX/PRX/DRX assignments and the measured OTA results are summarised in Table 9.7.

Table 9.7: *Measured TRP and TIS values for Configuration 3.*

Band	TRX Path	PRX/DRX Paths	TRP (dBm)	TIS (dBm)
n5/n8/n20	ANT0 (External)	DRX: ANT2 (SMD Main)	19.6	−77.4
n28	ANT3 (SMD Diversity)	PRX: ANT0 (External), DRX: ANT2 (SMD Main)	14.2	<i>Disconnected</i>

For the n5, n8, and n20 bands, a TRP of 19.6 dBm and a TIS of −77.4 dBm were obtained, as reported in Table 9.7. Although relocating the TRX path to the external antenna effectively reduced EMI coupling, the SMD main antenna assigned to the DRX path provided only limited benefit due to its relatively low radiation efficiency. Therefore, the observed performance indicates that the improvement in TIS was primarily driven by the TRX placement rather than diversity gain.

In contrast, for the n28 band, the TRX path was assigned to the SMD diversity antenna, which exhibited both low passive efficiency and partial EMI exposure. As a result, the link could not be stabilised, leading to reduced TRP (14.2 dBm) and a disconnected

TIS condition, as shown in Table 9.7. These findings confirm that even when diversity support is available, inadequate TRX antenna efficiency fundamentally limits downlink performance and cannot be compensated by PRX or DRX paths.

9.6.3.4 Configuration 4 — Final Four-Port Configuration

In the final configuration, three active antennas were combined to ensure simultaneous uplink and downlink stability across all tested bands. For the n5, n8, and n20 bands, the external antenna was assigned to the TRX path, while the SMD diversity antenna was connected to the DRX path. For the n28 band, the SMD main antenna was used for TRX, the external antenna for PRX, and the SMD diversity antenna for DRX. The corresponding port assignments and measurement results are summarised in Table 9.8.

Table 9.8: *Measured TRP and TIS values for Configuration 4.*

Band	TRX Path	PRX/DRX Paths	TRP (dBm)	TIS (dBm)
n5/n8/n20	ANT0 (External)	DRX: ANT2 (SMD Diversity)	19.7	−82.1
n28	ANT3 (SMD Main)	PRX: ANT0 (External), DRX: ANT2 (SMD Diversity)	20.8	−82.9

For the n5, n8, and n20 bands, this configuration resulted in a TRP of 19.7 dBm and a TIS of −82.1 dBm, as listed in Table 9.8. Compared to the previous configurations, the TRX relocation to the external antenna ensured reduced EMI interaction, while the DRX diversity branch provided additional downlink robustness.

For the n28 band, assigning the SMD main antenna to the TRX path provided sufficient transmit capability, while simultaneously using the external antenna as PRX and the SMD diversity antenna as DRX enabled a stable multi-branch receive structure. As a result, TRP of 20.8 dBm and TIS of −82.9 dBm were obtained, fully satisfying the performance requirements.

In this configuration, the TRX antenna selection primarily optimized uplink power,

whereas the external PRX and SMD diversity DRX paths significantly strengthened down-link reception. Consequently, stable connectivity was achieved across all measured channels, and this configuration satisfied the TRP and TIS targets for all bands. Therefore, Configuration 4 was selected as the final antenna layout, consistent with the decision process shown in Figure 9.11.



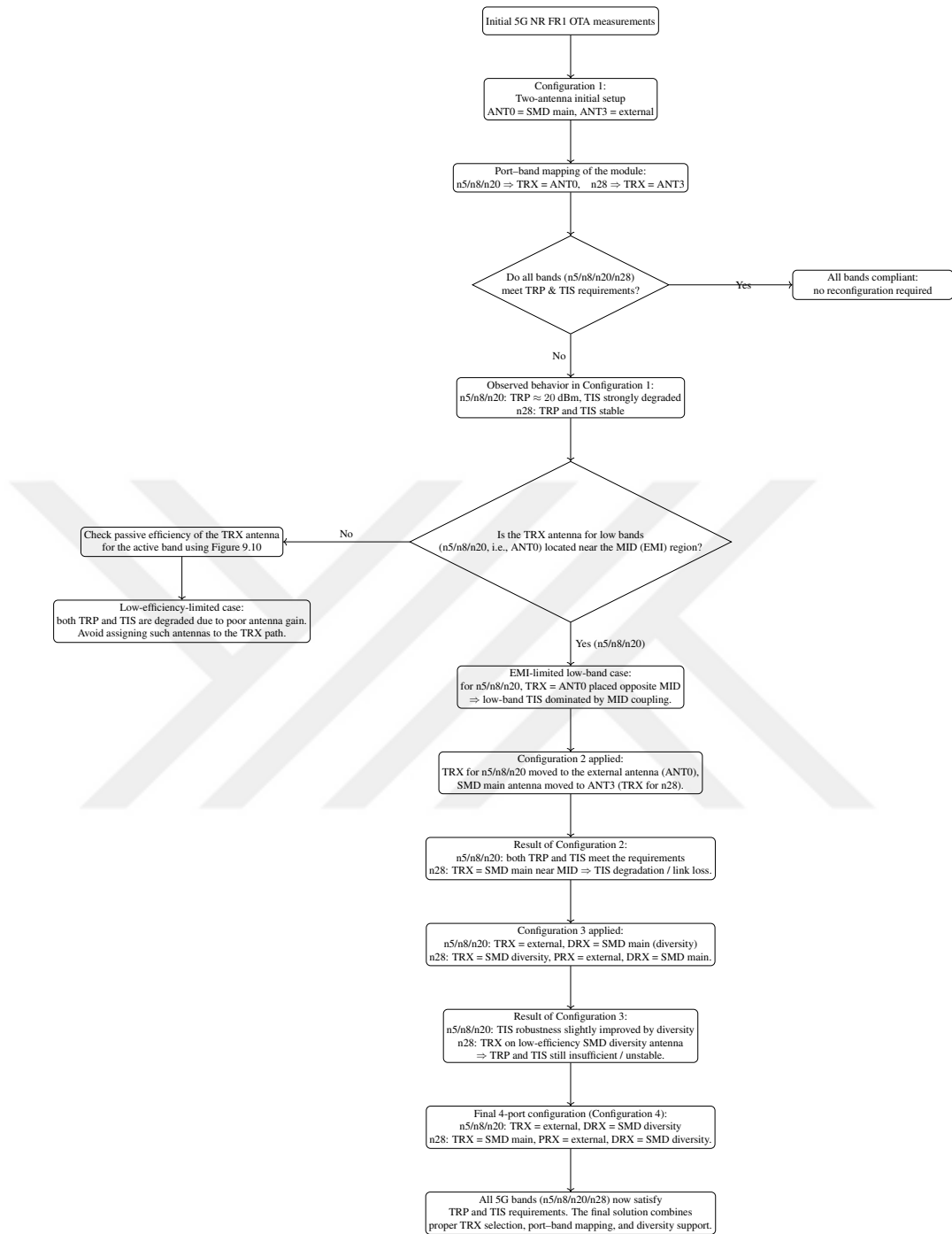


Figure 9.11: Band-aware engineering flow for 5G antenna reconfiguration, summarizing how port-band mapping, TRX location with respect to the MID region, antenna efficiency, and PRX/DRX diversity lead from the initial two-antenna setup (Configuration 1) to the final four-port solution (Configuration 4) that satisfies TRP and TIS requirements across n5, n8, n20, and n28 bands.

10. DISCUSSION AND FUTURE WORKS

Throughout this thesis, measurements and analyses have shown that the antenna performance degradation observed in AC EV chargers cannot be attributed to a single mechanism. The findings demonstrate that antenna design alone is not sufficient; mechanical layout, power distribution, near field interactions and electrically relevant structures inside the device jointly shape system level RF performance. Accordingly, the following points are considered important for similar designs.

The findings of this thesis collectively demonstrate that the interaction between antenna operating bands and the physical dimensions of the device plays a critical role. At low LTE bands, the long wavelength causes the antenna to behave electrically small, resulting in a large reactive near field region within the device. This thesis showed that PCB surfaces, the TFT LCD display module with its associated FPC cable, and power distribution paths located within this region can introduce electromagnetic coupling mechanisms that degrade performance. Therefore, evaluating which regions of the device volume fall inside the reactive near field for the targeted frequency bands is essential during design.

In addition, the results show that structures surrounding the antenna cannot be regarded as electromagnetically passive. The TFT LCD display unit, the FPC cable and power distribution traces located near the antenna can effectively become part of the antenna's electromagnetic environment, influencing impedance, current distribution or radiated power. For this reason, establishing coordination between mechanical design, power electronics and RF subsystems early in the design process provides significantly greater benefit than late-stage corrective actions. Independent development of these disciplines may lead to unexpected performance degradation once the system is assembled.

Furthermore, this thesis highlights the necessity of evaluating device size together with operating frequency. In compact devices such as mobile phones, the reactive near field can occupy the entire device volume, making both low and high bands sensitive to

structural interactions. In contrast, although AC EV chargers are physically large, measurements in this thesis showed that performance degradation emerged predominantly in the low bands. At low frequencies, long wavelengths make the antenna electrically small and expand the reactive region over a large volume, allowing nearby PCB structures, TFT LCD–FPC elements or power paths to fall within the field and limit TIS performance. At higher frequencies, however, the shorter wavelength results in a smaller reactive region, placing the same structures outside the antenna’s electromagnetic influence and preventing observable performance degradation. This finding indicates that a single antenna architecture may display completely different sensitivity characteristics across different bands.

Finally, this thesis confirms that effective corrective actions are product-dependent. The dominant limitation in the EVC01 platform was associated with PCB geometry, whereas in EVC08 the TFT LCD–FPC assembly was the primary contributor. In the 5G evaluation, antenna-to-port assignment became the determining factor. For this reason, proposing a single universally applicable solution is not appropriate. Instead, the recommended approach is to identify the dominant interaction mechanism at the system level and implement modifications targeting that mechanism. This represents a generalizable methodology: observe performance, identify the coupling path, and apply a focused solution.

These findings also provide opportunities for future work. Mapping the antenna’s reactive region within the enclosure, validating the influence of structures within this region through near field scanning or simulation, and studying different enclosure shapes and multi-antenna layouts may strengthen design guidance. Additionally, since this thesis evaluated only two platforms, verifying whether similar behaviors occur in other mechanical configurations remains an open research question.

Overall, this thesis demonstrates that antenna performance degradation in AC EV

chargers is governed not only by the antenna itself but by the electromagnetic environment of the entire device. These observations provide practical insight into which regions require attention during design and may help anticipate low band performance issues in future systems before physical testing.

As a future work, this thesis highlights a clear gap in the existing literature regarding the system level evaluation of antenna performance in AC EV chargers. Although an extensive research and testing background exists for compact consumer devices such as mobile phones and laptop computers, AC EV chargers despite their larger physical size have not been examined with comparable scope. Although the radio module and antenna satisfy specifications when evaluated individually, this thesis demonstrated that internal system interactions can lead to performance degradation, particularly across low-frequency bands. Therefore, repeating similar analyzes on different platforms, mechanical designs, and antenna integration strategies would be valuable for determining whether the mechanisms observed here are product specific or represent a more general design trend. Future work may also benefit from analyzing antenna PCB layouts and surrounding mechanical or electrical elements using full wave electromagnetic simulation tools such as High Frequency Structure Simulator (HFSS) or Computer Simulation Technology (CST) prior to prototyping. Evaluating field distributions and potential coupling regions in the design stage could help guide layout decisions and prevent the performance limitations observed in this thesis from appearing in final hardware. Finally, developing systematic design practices that ensure that no EMI sources (such as power traces, PCB structures, or TFT LCD panels) are placed within the antenna's reactive near field region is considered an important direction for proactively improving RF performance in future charger platforms.

11. CONCLUSIONS

In this thesis, the mechanisms limiting the performance of LTE and 5G communication in AC EV chargers have been systematically investigated. The results show that this problem cannot be attributed to a single cause, but arises from the combined influence of multiple physical interactions. The findings indicate that the RF performance degradation observed in AC EV chargers cannot be fully explained by classical EMI evaluations based only on radiation and conduction. Interaction between the antenna and the device, reactive near field behaviour, and local structural effects may also play a significant role, depending on the platform architecture. For this purpose, a measurement driven, theoretically supported diagnostic methodology has been applied to two EV charger platforms, namely EVC01 and EVC08. The methodology begins with full device OTA measurements, followed by subsystem isolation, analysis of the power distribution network and near field effects, identification of the dominant mechanisms through capacitive and common impedance coupling models, and finally verification of the proposed design improvements through repeated OTA measurements. This procedure enabled clear identification of whether the observed performance degradation originated from the antenna itself, the baseboard, or the complete system configuration.

The following subsections summarise the LTE performance of the EVC01 and EVC08 platforms, the 5G performance of the EVC08 platform, and the overall engineering and scientific contributions of this thesis.

Initial full device OTA measurements performed on the EVC01 platform revealed noticeable TIS degradation for LTE Band 28, Band 20, and Band 8. For instance, in the initial system configuration, the average TIS values were approximately -82.28 dBm for Band 20, -81.57 dBm for Band 8 and -80.46 dBm for Band 28, which are higher (less negative) than the commonly accepted -86 dBm requirement and therefore indicates insufficient receiver sensitivity, therefore falling short of the commonly accepted -86 dBm

requirement. Considering the ± 1.5 dBm measurement tolerance used in OTA testing, The practical acceptance threshold for TIS is taken as approximately -84.5 dBm, even though the nominal requirement is -86 dBm. Therefore, TIS values higher than (less negative than) -84.5 dBm are interpreted as insufficient receiver sensitivity.

Conducted sensitivity tests and OTA measurements under isolated baseboard conditions confirmed that the LTE module and the antenna, when evaluated independently from the system, met the expected performance levels. This clearly indicated that the performance degradation originated from electromagnetic interactions at the system level rather than an intrinsic limitation of the radio module. Experimental investigations demonstrated that PCB structures in the antenna's reactive near field significantly influence radiation behaviour. The ACCT board, being close to the antenna, altered the near field distribution. Removing the antenna facing portion of this PCB resulted in a measurable improvement in TIS performance. This verified that desensitization in low LTE bands on the EVC01 platform is strongly linked to reactive near field coupling.

In addition to geometric effects, the device's power distribution network played a decisive role. When the platform was supplied from the internal DC power network, TIS performance degraded. Supplying the device from an external DC source restored nominal performance. For example, on LTE Band 8, the TIS value was about -92 dBm when the baseboard was supplied externally, but it deteriorated to about -84 dBm with the internal SMPS network. This corresponds to a performance degradation of approximately 8 dBm. Inserting a CM filter into the internal power path improved TIS to about -87 dBm, meeting the -86 dBm requirement and confirming that the internal power network introduces additional noise via common impedance coupling.

In the final stage, both domains of improvement were combined: the PCB geometry was modified to reduce near field interaction, and the power distribution network was

improved using appropriate filtering. The full device OTA tests of this final configuration indicated that the mean TIS levels reached approximately -84.6 dBm for Band 20, -87.24 dBm for Band 8, and -85.3 dBm for Band 28. Compared to the initial results, this corresponds to an improvement in TIS of approximately 6 dBm for Band 8, an improvement of around 5 dBm for Band 28 and an improvement of approximately 3 dBm for Band 20, which brings all bands within the performance requirement when measurement tolerances are also considered. In summary, for the EVC01 platform it has been shown that RF performance is strongly linked to both antenna near field geometry and the power distribution network, and that the proposed methodology can effectively identify and mitigate the combined effects of reactive near field and common impedance coupling mechanisms.

In contrast to EVC01, the EVC08 platform features a more complex mechanical and electronic structure, including a main antenna, diversity antenna, TFT LCD panel, and MID unit. Here, initial LTE OTA measurements on this platform revealed that the TIS performance for Band 28 and Band 20, and both TIS and TRP performance for Band 8, were outside the required limits. For example, in the initial configuration the average TIS values were approximately -80.85 dBm for Band 28 and Band 20, and -83.9 dBm for Band 8, while Band 8 TRP was around 15.9 dBm. When the TFT LCD panel and its metallic rear enclosure were removed from the system, a significant improvement in RF performance was observed across all LTE bands. Under these conditions, TIS values reached approximately -88.4 dBm to -89.5 dBm for Band 28 and Band 20, around -90.9 dBm for Band 8, and Band 8 TRP increased to approximately 18.5 dBm. These observations confirmed that the panel structure and its metallic enclosure form a dominant EMI coupling element, which primarily affects the antenna through capacitive and resonance driven near field interactions.

Therefore, a capacitive coupling model was constructed to represent the panel

structure, the digital noise source, and the system ground. In this equivalent network, C_{12} represents the capacitive coupling between the digital source and the metallic panel, while C_{2G} models the capacitive coupling between the panel and the system ground. Due to weak grounding, the metallic panel exhibits a “floating conductive body’s behavior and accumulates unwanted voltage, feeding strong electric field components into the antenna near field region. The effective electrical size of the panel that becomes comparable to a fraction of the wavelength in the LTE Band further enhances the resonance conditions, making this band particularly sensitive.

Design improvements were therefore directly targeted at this physical coupling path. The high frequency impedance of the panel was reduced by establishing two low impedance grounding points, shifting the resonance behavior away from the LTE Band 8 region. In addition, the FPC cable connected to the display was mechanically stabilized and better referenced to the ground to suppress parasitic current paths. With this final configuration, full device OTA measurements were repeated, yielding average TIS levels of about -85.7 dBm for Band 28 and Band 20 and approximately -86.3 dBm for Band 8. Meanwhile, Band 8 TRP improved from roughly 15.9 dBm to approximately 21.0 dBm. Compared with the initial system measurements, these results correspond to an approximate TIS improvement of approximately $4\text{--}5$ dBm for Band 28 and Band 20, an improvement of approximately $2\text{--}3$ dBm for Band 8 TIS, and an improvement of approximately 5 dBm in Band 8 TRP. As a result, the TIS and TRP performance for all LTE bands returned to within the specification limits. These results demonstrate that in large embedded RF systems, panels, displays, and floating metallic structures can act as critical EMI elements and that properly defined grounding and stabilization strategies can restore RF performance without requiring excessive shielding or major structural redesign.

The 5G evaluation of the EVC08 platform exhibited a character distinctly different from its LTE performance. The system employs two SMD chip antennas and an external

FPC antenna together with the multi port RF architecture of the Quectel RM500U-EA 5G NR FR1 module. For the 5G case, particularly in bands such as n5/n8/n20, the primary limiting factor was not a classical limitation of EMI level; instead, the decisive factors were the placement of the TRX antenna relative to internal EMI regions, passive antenna efficiency in the final product, and fixed port–band assignment of the 5G module.

Passive characterization demonstrated that antenna performance in the real device differed significantly from that of ideal reference evaluation boards, highlighting the importance of product specific integration. The internal port–band mapping of the RM500U-EA was verified, confirming that certain 5G bands are bound to specific TRX/PRX/DRX ports.

In Configuration 1, the TRX path for bands n5/n8/n20 was assigned to the main SMD antenna located close to the EMI generating regions. In this case, TRP remained around 20.2 dBm, while TIS deteriorated to approximately -65.8 dBm. Meanwhile, n28 performance was stable because its TRX path was already assigned to the external antenna, producing around 20.3 dBm TRP and -81.1 dBm TIS.

In subsequent configurations, TRX and PRX/DRX assignments were varied. Moving the TRX path of the low bands to the external antenna increased the TIS to approximately -76.9 to -77.4 dBm; however, in some configurations the n28 receive path was disrupted, preventing stable measurement. These intermediate results showed that antenna diversity alone is not sufficient; the TRX antenna must simultaneously offer proper EMI separation and adequate passive efficiency.

In the final Configuration 4, a four port structure was designed. For n5/n8/n20, the TRX path was assigned to the external antenna and the DRX path to the SMD diversity antenna. For n28, the TRX path was assigned to the SMD main antenna, PRX to the external antenna, and DRX to the SMD diversity antenna. In this configuration, n5/n8/n20

achieved approximately 19.7 dBm TRP and -82.1 dBm TIS. while n28 achieved approximately 20.8 dBm TRP and -82.9 dBm TIS. Compared to Configuration 1, the TIS of the n5/n8/n20 group improved by approximately 16 to 17 dBm, highlighting the decisive role of TRX antenna placement and proper port–band mapping. These results confirm that, for the EVC08 5G platform, the dominant constraint is not panel induced EMI, but rather the combined effect of EMI distribution, antenna efficiency, and RF port assignment strategy. By aligning antenna placement with EMI behavior and properly configuring the RF port structure, stable and compliant 5G performance has been achieved.

Overall, this thesis demonstrates that RF performance degradation in AC electric vehicle chargers results from the combined influence of antenna integration, reactive near field behavior, floating conductive structures and the power distribution network, rather than a single dominant mechanism. A structured measurement driven methodology was established to identify dominant coupling mechanisms and guide effective design improvements. The application of this approach to two different charger platforms confirmed that correct addressing near field interaction, panel related coupling, and RF port configuration enables recovery of compliant LTE and 5G performance and ensures stable cellular communication.

In general, the findings highlight that AC EV chargers behave fundamentally differently from conventional handheld RF devices and must be analyzed with a more holistic electromagnetic perspective. The methodology and design insights presented in this thesis therefore offer a practical and physics based roadmap for improving RF robustness in complex embedded communication systems.

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