HYBRID METHODS FOR MIXED SIGNAL CIRCUITS
SUBJECT TO ON & OFF - BOARD
ELECTROMAGNETIC INTERFERENCE

DISSERTATION

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By

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* * * * *

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ABSTRACT

Electromagnetic Interference (EMI) disrupting operations of electronic systems has been studied for decades. Many theoretical techniques and protection mechanisms are proposed to reduce impact of EMI on electronic systems. However, a new breed of EMI, Intentional Electromagnetic Interference (IEMI) has recently gained significant attention due to its potential threat to computer networks and communication systems in addition to its potential applications by the military. In this dissertation, we carry out a theoretical and experimental study of mixed signal circuits subject to on & off – board IEMI. Our ultimate goal is to develop a general framework for the EMI analysis of complex systems comprised of cable bundles and Printed Circuit Boards (PCBs) housing mixed signal circuits. In this context, we first consider on–board EMI effects on digital circuits, particularly on an inverter to show the vulnerability of digital devices to RF interference and investigate both system and device level upsets due to adjacent EMI sources on PCBs. Next, we review port analysis techniques to show the applications of the S-Parameter matrix for on–board EMI/EMC analysis. Subsequently, we extent the port analysis method with hybrid S-parameters to account for external field coupling to mixed signal circuits. In other words, we introduce additional hybrid S-parameters that establish a link between the existing board ports and external EMI. Thus, we can handle both on – board and off – board EMI problems concurrently. The new hybrid S-Parameter
matrix can be integrated into standard circuit tools such as HSPICE and Advanced Design System (ADS, Agilent Technologies) and also allows for both time domain and Harmonic Balance simulations of non-linear RF-digital components via broadband network characterization. We lastly employ hybrid S-parameters to carry out experimental and theoretical studies on RF power amplifiers, performance evaluation of digital modulation schemes and digital timers to address external EMI on communication systems and automotive electronics.
Dedicated to those who deny the very existence of the box
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CHAPTER 1

INTRODUCTION

Electromagnetic Interference (EMI) analysis of electronic systems is becoming more important with the increased use of electronic systems, wireless communications, and recent security threats posed to electronic and computer systems. Experimental studies reported in [1], [2] have already shown that serious damage can be done to electronic systems with High Power Microwave (HPM) sources. The severity of such threats was emphasized with an experiment that hand-held HPM units, which can be located in a suitcase, can lead to serious damage on electronic systems from a reasonable distance [2]. In [3], it was also demonstrated that nonreversible upset could be made in fundamental semiconductor blocks such as logic and microcontroller units with Ultrawideband (UWB) HPM sources.

There are numerous unfortunate incidents in practice that are attributed to EMI. Our experience with USS Forrestal in 1967, which was one of the worst accidents due to EMI is a typical example of how high power Electromagnetic waves can lead to catastrophic accidents [4]. It was later found out that a military aircraft, during a carrier landing, was exposed to the ship’s radar. A degraded cable shield termination on the landing aircraft allowed coupling from the radar’s signal into the aircraft
causing accidental firing of its munitions hitting a fully armed and fueled aircraft on the deck. It resulted in 134 deaths.

Another fatal example of EMI was reported in the medical care industry [4]. An attached monitor and defibrillator in the ambulance turned off every time the radio transmitter was used. This caused the death of a 93 year old patient. Investigations concluded that it was primarily due to the fiberglass ambulance roof that permitted high levels of radiated fields inside the patient area of the ambulance.

We also have occurrences of EMI in automobiles. The most prominent example is when Antilock braking systems (ABS) was introduced. The cars in Germany on autobahn experienced EMI problems when they passed a nearby transmitter when brakes were applied [4]. This was easily resolved with replacement of a mesh screen around the pertinent electronic module controlling ABS.

Following the September 11th attacks, a special attention has been given to Intentional Electromagnetic Interference (IEMI) (see special issue on High Power Electromagnetic Waves and IEMI, IEEE Transactions on Electromagnetic Compatibility, August 2004). The primary motivation for the attention to EMI is very well summarized in [4] as follows

- Increased terrorist threats
- Increased proliferation of IEMI sources
- Increased dependence on information and automated safety and mission critical electronic systems
- Increased EM susceptibility of high-density IT systems operating at higher frequencies and lower voltages
Recent concerns in the finance sector, primarily in Wall Street, accentuates the fear of intentional EMI attack on financial computer and data processing systems. While some effort has been exerted towards protection and shielding of civil systems, there has been also military interest at IEMI which aim to upset enemy’s electronic systems [5].

We can categorize efforts to study IEMI on electronic systems in the following areas,

1. Understanding of the EMI phenomena and thus protect systems
2. Developing tools and weapons to upset electronic systems

Later in the chapter we discuss existing mitigation techniques to address the former. However, the latter is primarily of military interest and not addressed here.

In addition to increased efforts on the understanding of intentional EMI, recent advances in communication systems, specifically in wireless networks, lead to more concern on unintentional EMI and thus stipulating the understanding of different classes of EMI. Also, the number of antennas on automobiles rises (with increased use of satellite and land-communication systems), unintentional coupling to electronic systems becomes inevitable. Moreover, increased use of wireless communication at homes, electromagnetic compatibility and interference among electronic gadgets again necessitate understanding and analysis of unintentional EMI for better performance assessment. Examples of EMI concerns can be extended to commercial aircrafts and land mobile transportation systems.

Existing mitigation and protection techniques are primarily centered around enclosing and shielding circuit boards and cable bundles to minimize EMI effects. Not
only does this approach lead to high cost but also fails to perform as expected due to inadequate understanding of the coupling phenomena. A computational and experimental study presented in [6] points out that a small aperture on a metallic structure (possibly enclosing electronic systems) would be sufficient to couple high enough fields inside the enclosure at resonant frequencies to generate enough concern. Other studies led similar concerns as presented in [7], [8], [9], [10] and [11]. These studies suggest that metallic shielding of circuit boards may not be sufficient to protect electronic systems from external field coupling. Moreover, penetrating cable bundles play critical role in carrying EMI to the circuit terminals since they are more exposed to external field coupling. Therefore, shielding of cable bundles is critically important to protect electronic systems from EMI. However, such shielding practice becomes economically challenging in platforms such as automobiles and aircrafts where typical length of harness is in the order of tens of miles.

There are two types of EMI sources: (1) power sources attached to Printed Circuit Boards (PCBs) and cable bundles and (2) external RF waves, either intended or unintended. In this dissertation, we refer to the former as “on – board EMI sources”, whereas the latter shall be called as “off – board EMI source.”

Typical on-board EMI issues are listed as cross talk, signal integrity and radiation susceptibility. External field coupling from wireless communication systems and radars to electronic systems and cables are referred as off-board EMI issues.

The prevalent approach to on & off – board EMI evaluations is based on experimental studies. Measurements are typically carried out on a prototype of the test structure before mass production. However, experimental studies lead to inevitable increase in cost and time requirements on performance evaluation. Hence, theoretical
prediction of the EMI/EMC evaluations has become increasingly important. In the subsequent paragraphs, we discuss the state-of-the-art of theoretical analysis in an effort to place the work of this dissertation at an appropriate perspective.

1.1 State-of-the-Art for Theoretical Analysis of On & Off Board EMI on Cable Bundles and Mixed Signal Circuit Systems

Theoretical efforts for on & off-board EMI are centered around applying numerical tools such as the Finite Element Method (FEM), Method of Moments (MoM), Finite Difference Time Domain (FDTD) etc. for the enclosures, Multiconductor Transmission Line Theory for cable bundles and Modified Nodal Analysis (MNA) and Harmonic Balance Method (HBM) etc. for mixed signal circuit analysis. Below we describe recent developments of EMI analysis for cable bundles and mixed signal circuits.

1.1.1 EMI Analysis of Cable Bundles

Cable bundles constitute a large portion of many electronic systems. Because they are largely exposed to external threats and shielding efforts are provided at minimum level due to cost savings for commercial applications, they become an easy target to IEMI. Electromagnetic waves typically couple into circuits through cable bundles. Therefore, many studies have been undertaken for accurate characterization of field coupling to cable bundles. The most accurate prediction of EMI on harness would require full wave analysis of the test body with all the details and complexities included in the geometry. However, the geometrical complexities involved in cable bundles make such approach intractable to carry out realistic analysis. To date, the
primary approach on field coupling to cable bundles has been to apply Multiconductor Transmission Line Theory (MTLT), which modifies Telegrapher’s classical transmission line equations by incorporating additional voltage and current sources to model the external field excitation (off – board EMI) [12], [13], [14], [15], [16].

The advantage of MTLT is its capability to treat the test body and the harness separately. In other words, the incident field on the harness is computed in the absence of the cable bundles, thus enabling individual analysis of test body with full wave solvers. The computed incident field is next used as additional voltage and current sources within the MTLT formulation.

It should be however noted that MTLT is limited to cases where quasi-static conditions are met and yield fairly good results when the closest distance between the transmission line bundle and the surrounding ground structure is less than one tenth of the smallest wavelength [15]. In other words, MTLT yields accurate results for relatively low frequency analysis where the quasi-static approximation is known to yield reasonably good results [17], [18], [19].

MTLT does not account for self-radiation of the transmission lines (i.e. it neglects the common mode current). That is, MTLT can only predict the differential mode current, but for more complex surrounding structures, we also need to include the common mode current to better characterize the total current induced on the transmission lines in the high frequency regime. The common mode current is not an issue when the transmission lines are electrically very close to the surrounding ground structure. In this case, only the differential mode current is dominant and consequently the quasi-static approximation (low frequency analysis) is valid [20].
To improve the MTLT approximation (by accounting for radiation generated by the cable bundle), a technique that modifies Telegrapher’s coupling equations with additional iterative terms was proposed in [21] for a finite conductor over a perfectly conducting infinite plane. Also, Haase [22] proposed an iterative transmission line method, employing the Telegrapher’s equations. However, the technique was studied for non-uniform multiconductor transmission lines over a perfectly conducting flat plate.

Practical EMI analysis requires knowledge of the interactions among multiconductor transmission lines in presence of complex structures. To deal with such complexities, hybrid techniques (treating the complex structures with numerical methods and modeling the multiconductor transmission lines with MTLT) were considered in [18], [23], [24], [19], [25], [26], [27]. Electronic modules attached to the transmission lines can also be analyzed via equivalent SPICE models of the transmission lines extracted by employing the Partial Element Equivalent Circuit (PEEC) [28], [29], [30] under quasi-static conditions. However, as stated above, the classical MTLT approximation does not include the common mode current that is important in presence of such complex structures. In [23], an iterative approach was taken into account for the interactions between PCB traces and the surrounding structure. Nevertheless, the formulation proposed in [23] is based on an iterative solution of the quasi-static MTLT coupling equations. For PCBs, the quasi-static conditions are well justified, even in high frequencies and therefore; the standard MTLT method is adequate. Consequently, the approach in [23] is not suited for more complex structures as in the case with automobile harness.
In [27], we proposed a more general approach for handling complex cable bundles using SPICE to model conductors and Method of Moments for the surrounding structures. An iterative method, referred as Telegrapher’s Iterative Coupling Equations (TICE), was then used to handle the interaction between the surrounding structures and transmission line bundles. The key aspect of our formulation was that both the quasi-static and non-static effects were included. This was accomplished by introducing Telegrapher’s coupling equations derived in presence of complex structures. In this case, non-static terms were included via corrective iterative sources where the sources via the MTLT were updated subject to the additional field interactions from the surrounding structure. In other words, we proposed a general set of equations which were consistent with the physical insight of the phenomenon.

1.1.2 EMI Analysis of Mixed Signal Circuits

Even though the cable bundles play a critical role in transmitting the interfering signal to the circuit ports, response of the circuits and interaction between the circuit board and interfering signal dictate an accurate assessment of EMI. The challenge in EMI analysis of mixed signal circuits is to engage the governing equations - Kirchhoff’s Voltage (KVL) and Current Laws (KCL)- for the circuits with Maxwell’s equations for the EM structure. While the former is expressed in terms of voltages and currents, the latter is dictated by the equations in terms of electric and magnetic fields. Three different techniques are proposed in the literature to integrate KCL and KVL with Maxwell’s equations : (1) Transforming the entire EM structure into a circuit compatible form via the Partial Element Equivalent Circuit (PEEC) [31], [29], [30], [32], [33], [34] and solving the resulting lumped element
network via circuit solvers; (2) Employing port analysis via S-Parameters [35], [36], [37], [38], [39] to capture the EM effects and solving the resulting network via KVL and KCL; (3) Combining the Moment Method for modeling the EM structure with the Modified Modal Analysis for the circuits (MNA, [40]) [41] as a set of linear equations. In [41], generalized KVL for circuits with linear terminations were considered and definition of potential difference between two points inherently presumes quasi-static conditions. Furthermore, analysis in [41] also requires small signal equivalence (linear model) for non-linear devices since it solves linear system of equations for both EM structure and circuits. However, normal operation of many digital devices operate in non-linear regime and such practices may not be valid for susceptibility analysis in which large RF power may be injected to devices ports, as a result, forcing devices to operate in non-linear region. Therefore, PEEC and port analysis (S-parameter) are widely preferred for EMI/EMC analysis of mixed signal circuits. While the former can be employed for both on & off – board EMI problems in conjunction with Multiconductor Transmission Line Theory (MTLT) [42], [13], [15], [43], [44] the latter is used for handling only on – board EMI problems (until recently, see Fig. 1.1).

In addition to PEEC and port analysis, time domain techniques such as FDTD [45] and Time Domain Integral Equations (TDIE) [46] have been proposed for concurrent analysis of circuits and EM structures for both on & off – board EMI evaluations. Even though time domain full wave techniques yield reasonably accurate results, they suffer from computational inefficiencies due to the meshing of large volumes and simulation of RF circuits with high quality factors. Moreover, they may also run into convergence issues for circuit elements characterized with stiff differential equations.
and they are not practical for concurrent analysis with mixed time-frequency domain circuit solvers for RF applications such as Harmonic Balance Method.

Port analysis is more prevalent for EMI/EMC and signal integrity analysis of electronic designs due to its generality and practicality. Not only does it enable the analysis of PCBs with lumped linear circuit elements, it also allows handling of mixed RF-Digital circuits via broad-band characterization of the entire PCB. The first step in characterizing EM effects on a PCB is to model the entire board as an N-port S-parameter network. The computational cost associated with such an analysis is proportional to the number of ports assigned on the circuit board. However, we can increase its efficiency by only assigning ports for the non-linear elements since linear circuit components can be modeled in the EM analysis. Depending on the domain of the analysis (frequency or time domain), various techniques are proposed to integrate the resulting S-parameter network with circuit solvers such as SPICE for time domain ([47], [48], [49]) and Harmonic Balance (HBM) [50], [51], [52], [53] for mixed time-frequency domain analysis. A summary of state-of-the-art in EMI analysis of mixed signal circuits is displayed in Fig. 1.1.

1.2 State-of-the-Art in Device Level EMI Analysis of RF-Digital Systems: Experimental and Theoretical Studies

Many experimental studies have been primarily carried out in the literature focusing on single devices in presence of simple platforms lacking realistic complex surrounding structures [54], [55], [56], [57]. These studies present a set of observational data for the impact of EMI on device performance, but fails to capture understanding of the coupling phenomenon and its role in device performance. Furthermore, many experimental studies are not either coupled with theoretical analysis
1. Partial Element Equivalent Circuit (PEEC) for both On & Off Board EMI

Entire EM structure is transformed into a circuit compatible form via lumped elements

Advantages:
- Both time and frequency domain analysis
- Exploit numerical advances in circuit domain

Disadvantages:
- Numerically inefficient at high frequencies

2. {TDIE, FDTD, MoM} with MNA for both On & Off Board EMI

Combine EM analysis with Circuits in EM domain

Advantages:
- EM effects are fully captured

Disadvantages:
- Cannot exploit numerical advances customized for each domain
- Discretisation is dictated by the parameters in both domain

3. S-Parameters

Advantages:
- Freedom to choose the best solver for each domain
- Exploit numerical advances in each domain
- Suitable for optimization in both domain

Disadvantages:
- Numerically expensive for large number of ports
- Can handle only on-board EMI until recently

Figure 1.1: State-of-the-Art in EMI analysis of mixed signal circuits
that consequently leads to insufficient modeling and understanding of EMI on devices. Measurements in [1], [2] are representative of system level EMI in which systems in presence of complex realistic platforms subject to high power electromagnetic fields. However such measurements fail to explore what components in the system were disrupted and do not suggest theoretical analysis due to the complexities involved in the systems. In other words, there is a significant gap between experimental and the state-of-the-art theoretical tools in terms of characterizing and understanding device and system level EMI of complex platforms.

1.3 Objective of the Dissertation and Overview

Primary objective of this dissertation is to experimentally and theoretically analyze mixed signal circuits subject to electromagnetic interference. As accentuated in the previous sections, numerous computational EM and circuit tools have been developed making on-board EMI analysis of mixed signal circuits possible while external or susceptibility analysis were given little attention. Therefore, in this work, most of the focus is given to developing tools for off-board EMI analysis, primarily for intentional EMI. We base our theoretical analysis on the integration of EM tools with circuit solvers. We also conduct measurements to validate existing state-of-the-art tools and study device level failures in realistic platforms. This includes the EMI on digital modulation schemes and on electronic modules within automotive platforms.

More specifically, we perform experimental and theoretical on & off-board EMI analysis of an enclosure housing RF, analog and digital components. We experimentally carry out EMI studies on the fundamental circuit components such as a CMOS
inverter, an automotive timer, and an RF power amplifier in conjunction with digital modulation schemes.

1.3.1 Assumptions

Our theoretical analysis for off-board EMI primarily focuses on plane wave coupling in the upper range of the UHF Band (2-3 GHz). We also consider electronic circuits consist of both passive and active components (linear and nonlinear)-which simply include all mixed-signal circuits. Further, our analysis assumes that the Printed Circuit Board is comprised of multi-layer dielectric substrates with typical thicknesses on the order of 0.5-3.0 millimeters. Another assumption is that the physical absence of the circuit elements does not significantly alter the exact field distribution between the interconnects and the reference plane. This is a reasonable assumption considering that circuit components are very small compared to the smallest operational wavelength. Moreover, all conductors and substrates are assumed to be low loss in the frequency range of interest. Further, the primary transmission lines on the PCBs are presumed to be planar transmission lines, mostly microstrips. Even though the proposed analysis can be easily extended to lossy surfaces with small modification but we find it beyond the scope of this work given the limited resources and time to carry out this work.

In the initial phase of this research, we experimentally studied on-board EMI effects on digital circuits, particularly on an inverter to show the vulnerability of digital devices due to RF interference and to investigate both system and device level upsets due to adjacent EMI sources on Printed Circuit Boards. Next, we showed the validity and generality of S-Parameter method in evaluation of on-board EMI/EMC as well
as observed the similar upset mechanism with simulations. Details of this work is
given in the next chapter. Subsequently, recently developed hybrid scattering param-
eter method is explained and validation studies in conjunction with off – board EMI
analysis of the aforementioned inverter residing inside a metallic box with apertures
are given in chapter 3. Chapters 4 and 5 describe experimental and theoretical stud-
ies of the EMI effects on fundamental components of communication and automotive
electronics, respectively.
CHAPTER 2

ON – BOARD EMI ANALYSIS OF RF-DIGITAL SYSTEMS

Many on – board EMI problems such as cross talk and signal integrity have represented major challenges for EMC engineers to deal with. Identifying the potential issues before building the prototype requires highly comprehensive and complex tools that integrates circuit analysis with EM analysis. Therefore, there has been many studies in both circuit and EM domains to produce a computational method that would address practical EMI/EMC problems during the design stage. In this chapter, we review the prevalent techniques employed for on – board EMI analysis and carry out both experimental and theoretical analysis for an inverter subject to high power EMI. In evaluation of EMI effects on the performance of the inverter, we particularly focus on s-parameter technique that constitutes the foundations for the off – board EMI analysis described in the next Chapter. Conclusions drawn from the experimental analysis also serve as a basis for device and system level experimental studies given in Chapter 4 and 5.

In characterization of circuits, Kirchhoff’s current and voltage laws (KCL and KVL, respectively) have been employed for decades as the fundamental governing
equations in formulation in terms of nodal matrix and mixed or state-variable approaches [58]. Early efforts for computer-aided analysis of linear/nonlinear circuits date back to early 60s and accelerated in early 70s (TIME: [59], CANCER (Later SPICE): [60], ECAP: [61], BIAS-3: [62]). Many tools then experienced several hardware and theoretical limitations in terms of maximum number of nodes they could simulate, which was typically in the range of few hundreds. Further developments have led to a few standard packages currently in use. Among them are SPICE [63] and Harmonic Balance Simulator [50].

Early tools predominantly used nodal analysis and suffered in processing voltage sources and current-dependent circuit elements efficiently. With the introduction of Modified Nodal Analysis (MNA) [40], theoretical challenges in algebraic formulation of circuits were primarily tackled. MNA, based on KVL and KCL, still constitutes the core engine of SPICE and many other circuit tools today (see Fig. 2.1). Nevertheless, although these tools were very capable of characterization and analysis of circuit components, EM effects were highly neglected. To account for impact of transmission lines on PCBs, a novel approach, Partial Element Equivalent Circuit (PEEC) was introduced in [31]. PEEC characterized EM effects as inductors, capacitors and resistors. In other words, Maxwell’s equations were applied in a discretized manner and lumped element representation of EM components on the circuit board were obtained. The resulting lumped element network is combined with RF/circuit analysis under the same MNA matrix. For higher frequencies where retardation and radiation effects start to dominate, improvements are necessary to accommodate non-static EM effects [33], [34].
Kirchhoff’s Current Law (KCL):
Total current entering a node must be equal to the total current leaving the same node.

\[ I_1 + I_2 - I_3 - I_4 - I_5 = 0 \]

Kirchhoff’s Voltage Law (KVL):
Sum of the voltages in a closed loop is zero

**Linear Networks:**

\[ A\ddot{x}(t) + B\dot{x}(t) = c(t) \]

**Nonlinear Networks:**

\[ A\ddot{x}(t) + B\dot{x}(t) + F(x(t)) = c(t) \]

**Example Linear Network:**

\[ \begin{align*}
(\dot{V}_2 - \dot{V}_1)G_1 + I_0 &= -I_{f1} \\
-(\dot{V}_2 - \dot{V}_1)G_1 - V_2G_2 + (V_3 - V_2)G_3 &= I_{f1} \\
(V_3 - V_2)G_3 - CV_3 &= 0 \\
V_0 + V_2 - V_1 + 0 - V_2 &= 0
\end{align*} \]

\[ \begin{bmatrix}
0 & 0 & 0 & 0 & I_0 \\
0 & 0 & 0 & C & 0 \\
0 & 0 & 0 & 0 & I_0
\end{bmatrix}
\begin{bmatrix}
\dot{V}_1 \\
\dot{V}_2 \\
\dot{V}_3
\end{bmatrix}
\begin{bmatrix}
-G_1 & G_1 & 0 & 0 & 0 \\
G_1 & -G_1 - G_2 - G_1 & G_3 & 0 & 0 \\
0 & -G_3 & G_3 & 0 & 0
\end{bmatrix}
\begin{bmatrix}
V_1 \\
V_2 \\
V_3 \\
V_0 \\
I_{f1}
\end{bmatrix}
\]

\[ \dot{V}(t) = \frac{V(t) - V(t - \Delta t)}{\Delta t} \]

4 Unknowns
4 Equations

Figure 2.1: Summary of KVL, KCL and mathematical representation of linear/nonlinear circuits using an example network (upper right corner)
PEEC has been a core theory in many EM tools including Ansoft TPA® and Simlab PCBMod®. However, although it accommodates many EM effects on PCBs, its main challenge is the large MNA matrix for high frequency analysis. To overcome numerical burdensome associated with large MNA matrices, reduced order modeling methods have been considered in recent years [64], [65], [66]. These techniques enhance PEEC method and permits its use for electrically large and highly complicated PCBs at relatively ease. However, there are many challenges ahead to be addressed to make PEEC as practical as possible for highly complex large circuit networks.

An alternative approach to PEEC is port analysis (S-parameter) method that handles high frequency EM effects more efficiently and accurately for certain class of problems. Port analysis is a prevalent approach for EMI/EMC and signal integrity analysis of electronic designs due to its generality and practicality. In addition to analysis of PCBs with lumped linear circuit elements, it also allows for analysis of mixed signal circuits via broad-band characterization of the entire Printed Circuit Board. Depending on whether time or frequency domain analysis is considered, various techniques can be employed to integrate the S-parameter network with circuit solvers such as HSPICE for time domain [47], [48], [49] and Harmonic Balance Method [50], [51], [52], [53] for mixed time-frequency domain characterization.

There is no apparent advantage of PEEC over port analysis while both are capable of handling EMI problems. However, there is a prevalent consensus that in terms of computational efficiency and accuracy, one is more preferable to the other depending on how extensive EM structure is involved in the problem. For instance, as mentioned previously, at high frequencies, PEEC analysis results in a large MNA Matrix,
in other words increased number of lumped elements, due to the higher discretisation. Therefore, it can lead to intractable computational inefficiency both in EM and circuit analysis. In contrast, S-parameters are preferred for high frequency RF applications due to their ability to preserve EM effects more effectively and computationally more efficiently. They also provide freedom to choose the best solvers suited for the problem in EM and circuit domains. That also enables more efficient optimization studies since EM and circuit domains are taken into account independently. To state differently, one can choose the best solver for the EM analysis regardless what analysis technique is chosen for the circuit characterization. This implies analysis of large class of EMI/EMC problems involving mixed-signal circuit in presence of complex platforms. In this dissertation, we employ port modeling mainly due to the aforementioned advantages they offer. Below, we overview S-parameter method and its application in assessing impacts of EMI on the performance of an inverter we experimentally and theoretically studied.

2.1 Scattering Parameters and Port Analysis of On-Board EMI

S-parameters have been widely used in combining Electromagnetic analysis of transmission line networks with circuits involving of linear/nonlinear loads, especially for RF circuits. They have also been extensively studied for full wave extraction of parameters to characterize microwave structures [35], [67], [38], [68] and for integration of S-parameter networks with linear/nonlinear loads [69], [70], [49], [47], [71], [72], [48].

Port modeling is currently the core technology behind Ansoft Nexxim® for full wave analysis of RF-Digital systems.
As opposed to representing port relations in terms of voltage and currents via Z-parameters (Impedance) or Y-parameters (Admittance), S-parameters employ modal incident and reflected electromagnetic fields to establish a physics-based mathematical relation among the ports. To be specific, impedance and admittance parameters require ports to be terminated with open or short ports respectively. However, measurement of impedance and admittance parameters becomes impractical due to ambiguity of open and short ports in microwave frequencies. This drawback is easily circumvented with S-Parameter analysis, in which ports are terminated with a reference impedance, more prevalently 50Ω [71]. Because each mode has distinct characteristic impedance and propagation velocity, S-parameters are also referred to as modal parameters.

Needless to say, the frequency dependent nature of the modal waves and electrical size of structure require S-parameters to be computed for each frequency of interest. Therefore, dispersion characteristics of transmission lines are very well preserved within the S-matrix as well.

In computation of S-parameters, the reference impedance connected to the ports is assumed to be an infinitely long transmission line, which supports only the mode that S-parameters are computed for, to allow coupled evanescent modes decay along the reference line. However, quasi-TEM mode is strongly supported in microstrip lines for on-board voltage and current source excitations [73], [74], [75], [64], [76], [77] suggesting that evanescent and higher order modes can be ignored. It is also important to note that the typical ambiguity with voltage and current definition disappears for the quasi-TEM mode distribution [67] as well.
S-parameter method finds its applications mostly in analysis RF systems. In the subsequent section, we show that S-parameters can be readily employed for EMI analysis of digital systems as well. To this end, we discuss an example measurement conducted at Non-linear RF laboratory at The Ohio State University and discuss our theoretical analysis employing S-parameter technique.

2.2 Experimental and Computational On–Board EMI Characterization of an Inverter

We carried out on–board EMI measurements on a Texas Instruments (TI) inverter to observe system level upsets due to unmodulated (single tone), multi tone and pulse modulated interfering EM wave, as well as investigate the device level (soft) upset by injecting high power RF signal into the inverter input (see Fig. 2.2). We subsequently employ S-parameter approach for the theoretical characterization of the EMI effects observed in the measurement.
2.2.1 Measurement Setup

For measurements, we used a TI SN74AUC1GU04 inverter residing on a Duroid 5870 board with effective dielectric constant of 2.33 and thickness of 31 mils (See Fig. 2.3). The board was designed with Advanced Design System (ADS) of Agilent and manufactured with QuickCircuit in our facility at The Ohio State University. The EMI line is placed very close to the signal lines to establish a strong coupling to the inverter input and output ports. A Large Signal Network Analyzer (LSNA, see Fig. 2.4) was used to observe the incident and reflected waves at the inverter input and output ports. Our system is capable of measuring signals up to 20 harmonics for a signal with fundamental frequency of 1 GHz.

To generate a periodic pulse train for the inverter, the Agilent 81133A 3.35 GHz Pulse Generator was used. The pulse generator was connected to the input port of LSNA and synchronized with a 10 MHz reference line. LSNA works for periodic and periodically modulated signals. Therefore, it performs inverse fourier transform (IFFT) on the input pulses from the pulse generator with the user determining the number of the harmonics to retain. In this measurement, four harmonics of the
input signal were retained and shown to be adequate for characterizing the input signal. A computer was connected to LSNA for graphical display of the input/output data and in all measurements, the inverter was powered with 1.8V DC source. We used Anritsu MG3692A CW signal generator to couple EMI signal to EMI line in the board. Different modulations schemes were used as displayed in Fig. 2.5.

### 2.2.2 Port Analysis (S-Parameter Formulation) for the Simulations

Our goal for the simulations is to introduce and review with the S-parameter approach for the analysis of on–board EMI and to validate the approach before proceeding with the introduction of the hybrid S-parameter matrix for off–board EMI analysis in the next chapter. Port analysis was employed to capture the EM characteristics of the signal and EMI lines as displayed in Fig. 2.6. Since Texas Instruments
Figure 2.5: Setup describing EMI signals injected to the EMI line
Figure 2.6: Simulation setup for the computational characterization of the measurement

provides only an encrypted HSPICE model of the inverter, HSPICE of Synopsys was used for the simulations in conjunction with ADS for S-parameter modeling of the board. We characterized the board with 8-ports at the input, output and inverter terminals. The corresponding 8-Port S-Parameter network was obtained via ADS with respect to 50Ω reference impedance and exported to HSPICE. We remark that the bandwidth of the S-Parameter network was dictated by the time resolution used in the HSPICE analysis. To capture the rise and fall time characteristics, we specifically used time sampling of 80 ps (estimated from the inverter data sheet) within HSPICE. This corresponds to 6.25GHz \( f_{max} = \frac{0.5}{\Delta t} \), well above the 4\(^{th}\) harmonic of the inverter operating at 1 GHz. Thus, S-parameter network had sufficient bandwidth for our analysis.
2.2.3 Comparison of Measurement and Simulation Results

To investigate the EMI effects on the inverter, we carried out two different upset analysis: (1) system level upset and (2) device level upset. System level upset is mainly caused by EMI coupling to signal lines rather than a malfunction of the device. On the other hand, device level upset refers to soft (recoverable) upset of the device due to high RF power injection at the gates. For the former, we measure the signal at port P6, whereas port P5 is used for the latter as shown in Fig. 2.6.

Measurement and simulation results for the system level upset are given in Fig. 2.7a for an input pulse train at 0.6GHz with a 300 ps pulse width and 1.4V amplitude. The top right plot in this figure shows both the measurement and simulation results for an unperturbed signal(no EMI). Whereas, the bottom plot refers to the output signal when 25 dBm sinusoidal EMI signal at 2.4GHz and 3GHz is injected at port P1, respectively. Clearly, our simulations match the measurements quite well indicating the accuracy of the S-parameter port approach for EMI analysis of digital electronics.

We note that the inverter output was loaded with 50Ω well below the required typical loading on the order of 1KΩ. Therefore, the output remained quite below 1.8V. In addition to the low level output, it must be also noted that LSNA was not calibrated to capture the DC response. Therefore, the output is off and DC correction had to be applied to the simulations. We performed the similar analysis for the inverter operating at 1.0GHz and observed similar effects (see Fig. 2.7b). As in the previous case, our simulations are in very good agreement with measurement data.

For the case when a single tone EMI signal is turned on at port P1, the coupled EMI signal is superimposed on the output of the inverter displayed at the right bottom plot of Fig. 2.7b. The applied EMI input power was 25 dBm at 3 GHz corresponding
Figure 2.7: Inverter performance subject to single-tone EMI (a) System level upset: Inverter operating at 0.6GHz - (b) System level upset: Inverter operating at 1.0GHz
to the 3rd harmonic of the input periodic digital signal. Coupling to inverter input was computed to be approximately on the order of 15-20dB. This implies that about 5-10 dBm RF power at the input gates of the inverter will appear. However, the dominant interference is caused by direct coupling onto the output signal line rather than device malfunction. It is further obvious that the inverter performs the required function appropriately even though coupling from nearby EMI lines deteriorate the inverter input. This accentuates the importance of the cross-talk effects, which are similar to the interfering effects when the PCB is exposed to external EM wave.

We also considered impact of a narrow band EMI signals on the performance of the inverter. Fig. 2.8 displays inverter output for pulse modulated EMI (10 MHz bandwidth) injected into the inverter board, corresponding multi-tone EMI (10 MHz bandwidth) as displayed in Fig. 2.9. Description of pulse modulated and multi-tone EMI is given in Fig. 2.5. Inverter output is similar to the case of single tone EMI for both modulation types. Such observation implies that narrow band signals do exhibit as similar impacts on digital devices as single tone EMI. We also note that simulation of pulse modulated and multi-tone EMI setup would require HSPICERF for harmonic balance analysis. However, HSPICERF was not available. Therefore, we cannot carry out simulations verifying the measurements. Nevertheless, our previous single tone simulations demonstrates the validity of S-parameter modeling.

To cause device level upset, we increased the EMI power to 35 dBm and decreased the frequency down to 1 GHz corresponding to the dominant harmonic of the input digital pulse train. The output is measured right at the inverter output(port P6) and is plotted in the top plot of Fig. 2.10. We observe that since the coupling to the inverter input is not critically high, the inverter operates as expected. However,
Figure 2.8: Inverter performance subject to pulse modulated EMI (a) System level upset: Inverter operating at 0.6GHz - (b) System level upset: Inverter operating at 1.0GHz
Figure 2.9: Inverter performance subject to multi-tone EMI (a) System level upset: Inverter operating at 0.6GHz - (b) System level upset: Inverter operating at 1.0GHz
Figure 2.10: Device level upset: Inverter operating at 1.0GHz subject to high power EMI at 1.0GHz

when the EMI power is increased to 35 dBm (bottom plot in the same figure), we observe that high RF power on the order of 15 dBm is coupled to the inverter input and leading to decrease in output voltage level significantly. That is, even though the input signal is inverted, the output voltage level is small enough to consider the situation as device upset. The reason for the low output level can be attributed to the loss of saturation and the decrease in the transconductance of CMOS devices [78]. It is also seen that simulation results have higher voltage levels than measurements
when high power EMI is injected to inverter input. This is likely due to insufficient
SPICE characterization of the inverter for input signals above a certain range.
CHAPTER 3

HYBRID S-PARAMETERS FOR ON – OFF BOARD EMI ANALYSIS OF MIXED SIGNAL CIRCUITS

In the previous Chapter, we employed S-parameter method for validation of set of measurements on an inverter exposed to high power EMI. It was shown that S-parameter method was also highly efficient and accurate technique for handling on – board EMI analysis of digital devices. In this chapter, our objective is to extend S-parameter method to account for off – board EMI analysis of mixed-signal circuits in presence of complex structures.

External field coupling to PCBs and transmission line networks is primarily done with Multiconductor Transmission Line Theory [42], [13], [15], [43]. However, it is not suited well for accurate analysis at high frequencies since it inherently assumes strong quasi-static analysis in its formulation. Time domain techniques such as the Finite Difference Time Domain (FDTD) method have also been used for concurrent analysis of circuits and EM structures [45]. While time domain techniques also yield accurate results, they suffer from computational inefficiencies due to the meshing of large volumes and simulation of RF-devices with high quality factors. They also run into convergence problems for circuit elements with stiff differential equations.
As an alternative approach to the aforementioned methods to tackle their shortcomings, we have recently introduced hybrid S-parameter matrix that models transmission line networks with linear/nonlinear terminations subject to both plane wave and traditional port excitations [79], [80]. In this chapter, we extend the proposed analysis by decomposing external field excitation into forced and modal waves extracted via Generalized Pencil of Functions and present a more practical approach to integrate forced waves with circuit analysis tools such as Advanced Design System (ADS) and HSPICE. Proposed analysis also allows for integration of evanescent modes into circuit tools for high frequency analysis [67].

As opposed to traditional port excitations, external plane wave illumination leads to forced waves along the transmission line as well as propagating modal waves. Forced waves stem from enforcing phase matching with the incident wave along the structure walls and propagate with the wave number of the incident plane wave along the corresponding direction. Such forced waves are not affected by the loads attached to the ports. Conversely, backward and forward modal waves, originated from mismatches at port terminations, travel with corresponding eigenvalues that RF structure supports at the operating frequency. In our analysis, we consider forced waves as constant voltage sources at the ports and characterize the induced propagating modes with S-parameter matrix (hybrid S-parameters). The resulting S-matrix and voltage sources can then be exported to any circuit solver such as HSPICE and ADS and analyzed with the corresponding linear and non-linear port terminations. Since transmission line network is solely treated in EM domain and circuit components attached to the ports are handled in the circuit domain, numerical techniques customized for each domain can be fully exploited in our analysis. This approach
also allows for circuit design optimization without a need for repeated analysis of the microwave network.

Below, we first develop the theory for an arbitrary transmission line network subject to plane wave excitation. Next, we validate the proposed concept with a pair of transmission lines in free space excited by a plane wave with a current source attached to one of the terminals. Similarly, we extend our validation to a pair of microstrip lines on a PCB subject to plane wave excitation and more complex examples.

3.1 Theory

In this section, we consider characterization of the interactions among physical ports within a transmission line network using modal S-parameters. Subsequently, we propose a hybrid S-parameters matrix to include external plane wave excitations and proceed to describe techniques such as Generalized Pencil of Functions for the extraction of hybrid S-matrix entries. We then discuss how forced waves are treated at the ports for circuit analysis.

3.1.1 Modal S-Parameters for Coupling Among Physical Ports

Fig. 3.1 displays a typical mixed signal circuit board with nonuniform microstrip lines. For the characterization of such a board, we introduce the N-port S-parameter network giving (for the \( k^{th} \) mode),

\[
\begin{bmatrix}
    b^k_1 \\
    \vdots \\
    b^k_N
\end{bmatrix} = \begin{bmatrix}
    S^k_{1,1} & \cdots & S^k_{1,N} \\
    \vdots & \ddots & \vdots \\
    S^k_{N,1} & \cdots & S^k_{N,N}
\end{bmatrix}
\begin{bmatrix}
    a^k_1 \\
    \vdots \\
    a^k_N
\end{bmatrix}
\]

where \( S^k_{ij} = \frac{b^k_i}{a^k_j} \) with all ports terminated at their corresponding reference impedance \( Z_{ref_i} \), and \((a^k_i, b^k_i)\) referring to the incident and reflected waves, respectively, at the \( i^{th} \)
port. We must note that our analysis assume real reference impedance at the ports throughout our analysis. However, it can be readily extended to account for complex reference impedances at the ports with appropriate power relations [81], [82].

We remark that even though a PCB is shown in Fig. 3.1, our analysis applies to any transmission line configuration as is the case with multiconductor transmission lines and coaxial cable networks. As usual, the scattering matrix assumes the following field representation due to port excitation,

\[
\begin{align*}
E(s) &= \sum_k A_k e_k e^{-\gamma_k s} + \sum_k B_k e_k e^{\gamma_k s} \\
H(s) &= \sum_k C_k h_k e^{-\gamma_k s} + \sum_k D_k h_k e^{\gamma_k s}
\end{align*}
\]

where \(e_k\) and \(h_k\) refer to \(k^{th}\) modal electric and magnetic fields, respectively, with \(\gamma_k\) being the corresponding propagation constants whereas, \(A_k, B_k, C_k\) and \(D_k\) are the coefficients of the expansion.

In many publications, it has been shown that microstrip lines support hybrid modes (HE and EH) which are combination of TM and TE modes. In other words, pure TE and TM modes are not supported in microstrip lines due to the inhomogeneity in the cross sectional area of the structure [83], [84], [85], [86], [87], [75], [88]. EH mode denotes hybrid TE and TM modes with TE mode being dominant. Similarly, HE mode represents combination of TE and TM modes with TM mode being dominant. In [87], [89], [90], [91], [92], it was reported that for a typical PCB microstrip, the first three higher order modes are \(EH_1\), \(EH_2\) and \(EH_3\) in addition to the bound dominant propagating mode \(EH_0\) (quasi-TEM). \(EH_0\) mode corresponds to \(\gamma\) in (3.2) being pure imaginary and for higher order modes, \(\gamma\) has real part (attenuation/leakage), implying decay of the field in the longitudinal direction.
The above modal fields in (3.2) refer to the eigensolutions of the corresponding Sturm-Liouville problem subject to Dirichlet boundary conditions on PEC surfaces

\[ \hat{n} \times \mathbf{e}_k = 0 \]  
(3.3)

These modal fields must also satisfy the orthogonality condition

\[ \oint_S (\mathbf{e}_p \times \mathbf{h}_k) \cdot ds = \delta_{pk} \]  
(3.4)

with respect to the cross section of the propagation front along the transmission line.

To relate \((\mathbf{e}_k, \mathbf{h}_k)\) with terminal voltages and currents, we introduce the definitions

\[ V^k_i = - \int_{C_i} \overrightarrow{E_k} \cdot d\overrightarrow{l} \quad I^k_i = - \oint_{C_i} \overrightarrow{H_k} \cdot d\overrightarrow{l} \]  
(3.5)

where \((V^k, I^k)\) denote voltages and currents at the \(i^{th}\) port due to the \(k^{th}\) modal field. From (3.2), fields to be integrated are \(E_k = A_k \mathbf{e}_k e^{-\gamma_k s} + B_k \mathbf{e}_k e^{\gamma_k s}\) and \(H_k = C_k \mathbf{h}_k e^{-\gamma_k s} + D_k \mathbf{h}_k e^{\gamma_k s}\).

It should be noted that for a microstrip line, the total voltage varies with \(x\) and \(z\) where \((x, z)\) is the reference location on the microstrip line with respect to the ground.
plane. A typical approach to define a unique voltage is to average the voltage along the width of the microstrip line as follow [93], [94],

\[
V_{avg}^{total}(z) = \frac{1}{w} \int_{-w/2}^{w/2} V^{total}(x, z)dx = \sum_{m=1}^{n} (V_m^+ e^{-\gamma_m z} + V_m^- e^{\gamma_m z})
\]  

(3.6)

Mathematical relation between the \(k^{th}\) incident and reflected modal amplitudes, in (3.1) and modal voltages and currents in (3.5) is given by

\[
a_i^k = \frac{V_i^k + Z_{ref_i} I_i^k}{2\sqrt{Z_{ref_i}}} \quad b_i^k = \frac{V_i^k - Z_{ref_i} I_i^k}{2\sqrt{Z_{ref_i}}}
\]  

(3.7)

where \(Z_{ref_i}\) is the reference impedance for the corresponding \(i^{th}\) port. The expression in (3.7) can be construed as an interface between the circuit components (expressed in terms of voltages and currents) and the RF structure treated via full-wave electric and magnetic fields. However, we must note that (3.7) is only applicable to modal excitations at the physical ports. Therefore, one must account for non-modal field contributions at the ports for non-conventional port excitations. Below, we exploit the relation in (3.7) and introduce hybrid S-matrix approach to account for non-conventional external excitations such as plane waves on transmission line networks.

### 3.1.2 Hybrid S-Parameters for External Plane Wave Excitation

Let us consider the case of an external plane wave impinging upon an N-port arbitrary transmission line network shown in Fig. 3.3. Similar to the lumped port excitations, we propose to introduce the external plane wave as a source generated from an additional \((N + 1)^{th}\) port. We start our analysis by imposing Dirichlet
boundary conditions along the transmission line walls (see Fig. 3.2) as follows,

\[ \hat{n} \times \mathbf{E}_{total} = 0 \] (3.8)
\[ \hat{n} \times (\mathbf{E}_{inc} + \mathbf{E}_{scat}) = 0 \]

where \( \mathbf{E}_{inc} \) refers to electric field due to the incident plane wave in absence of the whole transmission line network, and \( \mathbf{E}_{scat} \) is the electric field radiated by the induced currents on the transmission line conductors.

For an infinitely long transmission line, (3.8) implies that \( \hat{n} \times \mathbf{E}_{scat} = -\hat{n} \times \mathbf{E}_{inc} \) at transmission line surfaces. However, for a finite transmission line, the reflected currents at the terminals will lead to modal fields which already satisfy the boundary conditions given in (3.3). Therefore, the scattered fields at the transmission line surfaces satisfy the conditions

\[ \hat{n} \times \mathbf{E}_{scat} = \hat{n} \times (-\mathbf{E}_{inc} + \mathbf{E}_{modal}) \] (3.9)
where $E_{\text{modal}}$ is given by (3.2) with the boundary conditions along the transmission line surfaces,

$$\hat{n} \times E_{\text{modal}} = 0$$  \hspace{1cm} (3.10)

Referring to (3.9) and [95], we observe that plane wave incidence on a transmission line introduces a forced wave (having the same wave number as the incident field) in addition to the modal fields. Thus, in the case of a plane wave excitation, we introduce the representation

$$E_{\text{total}} = E_{\text{forced}} + \sum_k A_k e_k e^{-\gamma_k s} + \sum_k B_k e_k e^{\gamma_k s}$$

$$E_{\text{total}} = E_{\text{forced}} + E_{\text{modal}}$$  \hspace{1cm} (3.11)

Comparing the above expression with (3.2), we deduce that the difference between the plane wave and lumped port excitation is that the former induces forced waves in addition to modal fields.

To account for the modal waves coupled to the physical ports at the transmission line terminals, we treat the plane wave as an additional $(N+1)^{th}$ port and modify the existing $N$-port S-matrix accordingly for each modal wave,

$$\begin{bmatrix}
  b^k_1 \\
  \vdots \\
  b^k_N \\
  b^k_{N+1}
\end{bmatrix} = \begin{bmatrix}
  S^k_{1,1} & \cdots & S^k_{1,N} & HS^k_{1,N+1} \\
  \vdots & \ddots & \vdots & \vdots \\
  S^k_{N,1} & \cdots & S^k_{N,N} & HS^k_{N,N+1} \\
  S^k_{N+1,1} & \cdots & S^k_{N+1,N} & HS^k_{N+1,N+1}
\end{bmatrix} \begin{bmatrix}
  a^k_1 \\
  \vdots \\
  a^k_N \\
  a^k_{N+1}
\end{bmatrix}$$

As seen, the $(N+1)^{th}$ port is characterized by the hybrid S-parameters, $HS^k_{i,N+1}$, representing plane wave coupling to $i^{th}$ port for the $k^{th}$ excited mode. Fig. 3.3 clearly demonstrates that plane wave is included as additional port in the circuit domain and
we treat forced waves as additional constant voltages at the ports. In the subsequent sections, we first describe extraction of hybrid S-parameters. Next, we explain how we integrate forced waves with circuit analysis.

3.1.3 Hybrid S-Parameters via Open Circuit Analysis

To calculate the hybrid S-parameters, we exploit the inherent relation between the incident and reflected waves, and voltage and currents at the ports given in (3.7). We first introduce the corresponding hybrid impedance matrix for the (N+1)-port network representing voltage and current relations at the ports due to modal and plane wave excitations,

\[
\begin{bmatrix}
V_1 \\
\vdots \\
V_N \\
\end{bmatrix}
= 
\begin{bmatrix}
Z_{1,1} & \cdots & Z_{1,N} & HZ_{1,N+1} \\
\vdots & \ddots & \vdots & \vdots \\
Z_{N,1} & \cdots & Z_{N,N} & HZ_{N,N+1} \\
\end{bmatrix}
\begin{bmatrix}
I_1 \\
\vdots \\
I_N \\
a_{N+1} \\
\end{bmatrix}
= [Z] \{\bar{I}\} + \{V_{oc}^{modal}\} \tag{3.12}
\]

where \(\{V_{oc}^{modal}\} = \{HZ\}a_{N+1}\) refers to open circuit voltage at the ports due to modal fields excited by the external plane wave. Coupling to the \((N+1)^{th}\) port is of no interest in our analysis. Therefore, it is excluded in the Z-matrix (namely, the \((N+1)^{th}\) row of the Z-matrix is set to zero). This also helps us circumvent mismatch problems at the EMI port. The last element added to the \(\{\bar{I}\}\) column, \(a_{N+1}\), represents the normalized plane wave coefficient. Thus, the column, \(\{HZ\}\) can be construed as that relating the open circuit modal voltages at the ports to the incident plane wave excitation.

To associate impedance matrix entries with S-parameters, we employ (3.7) to update (3.12) giving

\[
\sqrt{[Z_{ref}]}(\{\bar{a}\} + \{\bar{b}\}) = [Z](\sqrt{[Z_{ref}]}^{-1})(\{\bar{a}\} - \{\bar{b}\}) + \{V_{oc}^{modal}\} \tag{3.13}
\]
Figure 3.3: (a) Typical mixed-signal PCB subject to plane wave excitation (b) Circuit representation of (a) via hybrid port modeling of plane wave
where

\[
\{a\}^T = \{a_1 \cdots a_N\} \quad \{b\}^T = \{b_1 \cdots b_N\} \quad (3.14)
\]

\[
[Z] = \begin{bmatrix}
Z_{1,1} & \cdots & Z_{1,N} \\
\vdots & \ddots & \vdots \\
Z_{N,1} & \cdots & Z_{N,N}
\end{bmatrix}
\sqrt{[Z_{\text{ref}}]} = \begin{bmatrix}
\sqrt{Z_{\text{ref}_1}} & \cdots \\
\vdots \\
\sqrt{Z_{\text{ref}_N}}
\end{bmatrix}
\]

in which \([Z]\) and \(\sqrt{[Z_{\text{ref}}]}\) are already known matrices.

Rearranging the terms, we find that the coefficients of the incident and reflected waves at the physical ports are given by

\[
\{\bar{b}\} = \{[Z](\sqrt{[Z_{\text{ref}}]})^{-1} + \sqrt{[Z_{\text{ref}}]}\}^{-1}\{[Z](\sqrt{[Z_{\text{ref}}]})^{-1} - \sqrt{[Z_{\text{ref}}]}\}\{\bar{a}\}
\]

\[
+\{[Z](\sqrt{[Z_{\text{ref}}]})^{-1} + \sqrt{[Z_{\text{ref}}]}\}^{-1}\{V_{\text{modal}}\}
\]

Comparing (3.15) with (3.12) and setting \(a_{N+1} = \frac{|E_0|}{\sqrt{Z_{\text{ref}_{N+1}}}}\) to normalize the incident plane wave, we readily identify that

\[
\{\overline{H}S\} = \frac{\sqrt{Z_{\text{ref}_{N+1}}}}{|E_0|} \{[Z](\sqrt{[Z_{\text{ref}}]})^{-1} + \sqrt{[Z_{\text{ref}}]}\}^{-1}\{V_{\text{modal}}\}
\]

where \(|E_0|\) is the magnitude of the incident plane wave and \(\{\overline{H}S\}^T = \{HS_1 \cdots HS_N\}\).

The evaluation of \([Z_{ij}]\) in (3.12) is done in the usual manner via open circuit analysis. However, the evaluation of \(V_{\text{modal}}\) requires more attention. Once the open circuit modal voltages are obtained, \(\{\overline{H}S\}\) can be calculated via (3.16). Since the forced voltages do not depend on the terminations, they can be directly exported to the circuit solver shown in Fig. 3.3.

### 3.1.4 Generalized Pencil of Functions for Extraction of Hybrid S-Parameters

As noted above, the hybrid scattering matrix assumes the propagation of a discrete set of modes within the network. Knowledge of these modes and their associated parameters (e.g. \(\gamma_k, A_k\) and \(B_k\) as in (3.2)) is necessary for the extraction
of the hybrid S-matrix entries. Generalized Pencil of Functions [96], [97] method can be employed for the extraction of these parameters. Such an analysis has been successfully employed in the literature [89], [98]. Specifically, in [89] and [98], the current induced on a microstrip line is decomposed into the bound (dominant) and higher order modes and authors employed Generalized Pencil of Functions to find the corresponding mode amplitude and propagation constants to achieve the best fit. Similarly, in [99], FDTD was used in conjunction with Generalized Pencil of Functions to extract the S-Parameters of a waveguide structure via a full wave analysis. Further, in [100], Generalized Pencil of Functions was used to extract the parameters of current induced on large scatterers represented with sum of complex exponentials.

Once the parameters of the exponential terms in (3.11) are attained, one can readily distinguish forced and dominant modal waves by examining the propagation constants such that the dominant modal terms appear as a pair of backward and forward traveling voltages with negligible decay/attenuation constant. A more rigorous comparison can be also made by computing the propagation constants of the transmission line network by invoking the eigenfunction representation with the appropriate boundary conditions. Since the computed eigenvalues correspond to the propagation constants, one can then extract the modal propagation constants from Generalized Pencil of Functions Method results.

3.1.5 Integration of Forced Waves with Circuit Analysis

As described above, forced and modal waves can be extracted via Generalized Pencil of Functions Method. Also, we have shown that modal waves can be combined
with circuit analysis through the hybrid S-parameter matrix. Next, we describe incorporation of forced waves into the circuit analysis.

We start our analysis with the following impedance boundary condition that must be satisfied regardless of linear/nonlinear loads attached to the ports. Specifically, we have

\[ \mathbf{E}_{\text{tan}} = Z_s \mathbf{H}_{\text{tan}} \]  

(3.16)

where \( Z_s \) is measured in \( \Omega \) per square unit cell. The impedance boundary conditions for EM analysis at the ports can be translated into Ohm’s law to relate the voltage and currents at the terminals using the general form,

\[ V_{\text{total}} = f(Z_L, I_{\text{total}}) \]  

(3.17)

where \( Z_L \) is the complex linear/nonlinear load impedance at the ports. We must remark that the surface impedance \( Z_s \) in (3.5) can be expressed in terms of the total port impedance \( Z_L \) and the port dimensions. For instance, such a relation for the microstrip lines can be written as

\[ Z_L = Z_s \frac{h}{w} \]  

(3.18)

where \( h \) and \( w \) correspond to the height of the transmission line from the ground plane and the width of the strip line, respectively.

For plane wave excitation, the total voltage at the port is expressed in terms of modal and forced waves via (3.11) and (3.5), viz.

\[ V_{\text{total}} = V_{\text{forced}} + V_{\text{modal}} \]  

(3.19)

\[ V_{\text{modal}} = V_{\text{total}} - V_{\text{forced}} \]
where $V_{total}$ represents the total voltage at the terminals of any linear or nonlinear loads. As stated, $V_{forced}$ is a constant term, not associated with the loads at the ports. Thus, the forced voltage can be added to the ports as a constant source term to enforce Ohm’s law in circuit domain or equivalently the surface impedance boundary condition (see Fig. 3.3).

3.2 Validation Studies

To demonstrate the validity of the hybrid S-parameters, we first consider a pair of transmission lines subject to concurrent plane wave and a direct port excitation. Subsequently, we consider a more complex configuration consisting of a pair of microstrip lines on a PCB illuminated by an obliquely incident plane wave.

3.2.1 A Pair of Transmission Lines Subject to Plane Wave Excitation

Consider the Transmission Lines (TL) shown in Fig. 3.4 excited by a current source at the left and terminated by a load $Z_L = 100$ located 250mm from the source. The TL is comprised of two wires of radius 0.125mm and separated by 2mm having a characteristic impedance of $Z_0 = 332.24\,\Omega$ [44]. We are interested in computing the
voltage induced at the load when the TL is concurrently illuminated by a plane wave operating at 2 GHz (the same as the port source). This problem is therefore a typical EMI/EMC coupling analysis.

We consider the plane wave excitation \( \vec{E} = \vec{E}_{\text{inc}} e^{-j\vec{k} \cdot \vec{r}} \) with \( \vec{E}_{\text{inc}} = (\hat{x}1000 + \hat{y}500 + \hat{z}1500) \text{V/m} \) and \( \vec{k} = k_0 (\hat{x} + \hat{y} - \hat{z}) / \sqrt{3} \). Further, we assume that the current only flows in the \( \hat{y} \) direction since the wire radius is much smaller than the wavelength at the operating frequency.

We break down our analysis into two sections

1. Current source excitation
2. Plane wave excitation

Such an approach implicitly assumes linear circuit components attached to the ports. However, we must note that the proposed solution can be applied to the cases where non-linear loads are included by employing broad band S-parameter characterization.

**Current Source Excitation**

To compute the total voltage induced at the load, we employ S-parameter matrix defined for two ports where the current source and lumped ports are attached, respectively.

Since the current source supports quasi-TEM modes along the transmission line, one can establish a 2-Port S-Parameter network based on a quasi-TEM mode propagation. The resulting 2-Port S-Parameter network can be exported to any circuit simulator such as ADS (see Fig. 3.5) and connect the current source and the load at the corresponding ports.
After performing a full wave analysis (HFSS), we extracted the $2 \times 2$ S-parameter matrix

$$S_{2 \times 2} = \begin{bmatrix} 0.909 \angle 8.5 & 0.415 \angle 97.7 \\ 0.415 \angle 97.7 & 0.909 \angle 8.5 \end{bmatrix}$$ (3.20)

Subsequently, we exported the resulting S-matrix to ADS with the connected current source at port 1 and the load at port 2. Using ADS (see Fig. 3.5) we can then find the load voltage as a function of the current source.

![2-Port circuit representation of the transmission lines](image)

**Figure 3.5: 2-Port circuit representation of the transmission lines**

Table 3.1 shows a comparison of the full wave results with the proposed S-matrix/ADS simulation. As seen, an excellent agreement is achieved.

<table>
<thead>
<tr>
<th>I=10mA and $Z_L = 100\Omega$</th>
<th>Mag(VL)(Full Wave-HFSS)</th>
<th>Mag(VL) (2-Port Network-ADS)</th>
<th>Angle(VL)(Full Wave-HFSS)</th>
<th>Angle(VL)(2-Port Network-ADS)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1.87</td>
<td>1.89</td>
<td>2.31</td>
<td>2.51</td>
</tr>
</tbody>
</table>

Table 3.1: Voltage induced at the $Z_L = 100\Omega$ due to a current source of 10mA

1For the full wave analysis, we further use strips in place of wires by employing the standard equivalence $a = \frac{w}{4}$ where the $a$ is radius of the wire and $w$ is the width of the equivalent strip.
**Plane Wave Excitation**

We now proceed to include the plane wave coupling in terms of traveling wave components [95]. Referring to Fig. 3.6, the $C_1$ and $D_1$ coefficients correspond to coupling onto infinite transmission line and have the same phase as the incident field to force phase matching along the wires. The remaining terms represent forward and backward travelling (modal) currents and the total voltage along the TL can be also expressed as,

$$V_{total} = V_{forced} + V_{modal}$$

$$V_{total} = V_{inc}e^{-j \frac{k_0 y}{\sqrt{3}}} + V_1e^{-j \frac{k_0 y}{\sqrt{3}}} + V_2e^{jk_0 y} + V_3e^{-jk_0 y}$$

$$V_{forced} = (V_{inc} + V_1)e^{-j \frac{k_0 y}{\sqrt{3}}}$$

$$V_{modal} = V_2e^{jk_0 y} + V_3e^{-jk_0 y}$$

where $V_{inc}$ is the voltage due to the plane wave in the absence of the wires and $V_1$, $V_2$, and $V_3$ are associated with the induced currents $(C_1, D_1)$, $(C_2, D_2)$ and $(C_3, D_3)$, respectively.

The $V_{inc}$ and $V_1$ terms depend only on the polarization and magnitude of the plane wave excitation and independent of the loads connected to the wire terminals. Therefore, they are forced waves and independent of the attached loads. Further, their presence disappears at the moment the incident field stops to exist, even the modal fields continue their presence.

To substantiate the above expansion, we carried out a full wave analysis of the TL in Fig. 3.6 with plane wave excitation and extracted the resulting travelling wave representation via the Generalized Pencil of Functions method. Two sets of loads $(ZL_1, ZL_2)$ were used at each port.
Figure 3.6: Physical current decomposition of the modes traveling along the transmission line

For the loads $ZL_1 = 250\,\Omega$, $ZL_2 = 800\,\Omega$, we found that

$$V_{scat_1} = (0.57 + j0.04)e^{-j24.69y} + (-0.37 - j1.29)e^{+j42.76y}$$

$$+(-0.74 - j0.61)e^{-j42.03y}$$

and for the loads $ZL_1 = 400\,\Omega$, $ZL_2 = 400\,\Omega$, we obtained

$$V_{scat_1} = (0.56 + j0.05)e^{-j24.46y} + (-0.77 - j1.29)e^{+j42.77y}$$

$$+(-0.16 - j0.37)e^{-j41.84y}$$

Considering that $k_0 = \frac{2\pi}{\lambda} = \frac{2\pi}{0.15} = 41.88$ and $\frac{k_0}{\sqrt{3}} = 24.14$, it is clear that (3.22) and (3.23) are in agreement with (3.21). In other words, forced voltage terms did not alter with changing loads while modal waves responded to the attached loads. Based on this claim, one can establish a hybrid S-parameters network based on the quasi-TEM travelling voltages ($V_2$ and $V_3$) in conjunction with the 2-port S-parameter network constructed in the previous section. Similarly, the forced voltage terms can be included in the analysis as constant voltage sources at the ports.
3.2.2 Hybrid 3-Port Quasi-TEM S-Parameter Network Construction

In this section, we proceeded to construct a 3-Port Hybrid S-Parameter network such that Port 1 and Port 2 are physical ports at the terminals of the transmission line with Port-3 representing the plane wave source leading to the quasi-TEM wave induced along the transmission lines. We first computed the open circuit modal voltages at the ports and employed (3.16) to compute the hybrid S-parameters. The resulting $3 \times 3$ hybrid S-matrix is

$$S_{3 \times 3} = \begin{bmatrix}
0.909 \angle 8.5 & 0.415 \angle 97.7 & 0.187 \angle -11.81 \\
0.415 \angle 97.7 & 0.909 \angle 8.5 & 0.190 \angle 170.0 \\
0.0 \angle 0.0 & 0.0 \angle 0.0 & 0.0 \angle 0.0
\end{bmatrix}$$ (3.24)

where we set the last row to zero because only $HS_{1,3}$ and $HS_{2,3}$ are non-zero since they represent the coupling of the incident plane wave onto the physical ports at the transmission line terminals.

Next, we proceeded to employ ADS in conjunction with (3.24) to find the port voltages (see Fig. 3.7).

We performed three studies in which current source is set to zero, 5mA and 10mA respectively. Next, we compared the proposed method solution with full wave solution for the voltage induced at the load $Z_L$ for each current source and plane wave excitation. It is clearly observed in Fig. 3.8 that hybrid S-parameters agree very well with full wave results.
3.2.3 A Pair of Coupled Microstrip Lines Subject To Concurrent Plane Wave and On-Board Current Source Excitation

We now consider the geometry in Fig. 3.9, displaying a pair of coupled microstrip lines residing on a RT/Duroid 5880 board with 2.2 dielectric constant and thickness of 31mils (0.7874mm).

<table>
<thead>
<tr>
<th>Port</th>
<th>Termination</th>
</tr>
</thead>
<tbody>
<tr>
<td>P1</td>
<td>$I=50\text{mA}$</td>
</tr>
<tr>
<td>P2</td>
<td>$Z_2 = 10 - j250\Omega$</td>
</tr>
<tr>
<td>P3</td>
<td>$Z_3 = 100 - j100\Omega$</td>
</tr>
<tr>
<td>P4</td>
<td>$Z_4 = 400 + j50\Omega$</td>
</tr>
</tbody>
</table>

Table 3.2: Port terminations in Fig. 3.9
Figure 3.8: Comparison of hybrid S-parameter method results with full wave results for the validation problem Fig. 3.4 - (a) Amplitude of the voltage induced at the load $Z_L = 100\Omega$ for various current source values - (b) Phase of the voltage induced at the load $Z_L = 100\Omega$ for various current source values

The microstrip lines are terminated with complex impedances and an on-board current source was placed in port 1 (P1) given in Table 3.2. Additionally, a plane wave operating at 2.5GHz (the same as the current source) also impinged on the microstrip lines.

$$\vec{E} = \overline{E}_{inc} e^{-j\vec{k}.\vec{r}} \quad \text{with} \quad \overline{E}_{inc} = (\hat{x}2000 + \hat{y}2000 + \hat{z}3000)V/m \quad \text{and} \quad \vec{k} = k_0 \frac{20.5 + \hat{y} - \hat{z}}{1.5}$$

In our analysis, we aim to find the total voltage induced at the ports. To do so, we first extracted the standard $4 \times 4$ S-parameter matrix for the given port configuration with respect to $50\Omega$ reference impedance. Afterwards, we performed full wave analysis on the structure (with open ports) subject to only plane wave
excitation. Subsequently, we conducted Generalized Pencil of Functions analysis to extract the forced voltage and propagating modes along the lines. Upon obtaining modal and forced voltages at the ports, we then employed (3.16) to extract the hybrid S-parameters. Resulting $5 \times 5$ hybrid S-parameter matrix is exported to ADS and corresponding forced voltages and port terminations are connected to the respective ports (see Fig. 3.10).

![Figure 3.9: A pair of coupled microstrip lines residing on a PCB and subject to oblique plane wave illumination](image)

![Figure 3.10: Circuit representation of the coupled microstrip lines subject to concurrent plane wave and current source excitation](image)
In this configuration, ports $P1 - P4$ correspond to physical ports at the terminals of the coupled microstrip lines and port $P5$ represents the plane wave coupling. Circuit analysis was run at 2.5 GHz and performance of the proposed method is compared with full wave results (see Fig. 3.11). It is clearly observed that proposed method results agree well with full wave results.

Figure 3.11: Comparison of the proposed method with full wave results for the voltage induced at the ports shown in Fig. 3.9 - (a) Total induced voltage amplitude at each port - (b) Total induced voltage phase at each port

3.2.4 A More Complex Pair of Transmission Lines Subject to External EMI

We continue our validation with a more complex pair of microstrip lines residing on a RT/Duroid 5870 board subject to a plane wave illumination operating at 2 GHz (see Fig. 3.12). Electric field is E-polarized with $E_x = 1000 V/m$ and $E_y = 500 V/m$ and incident with $\hat{k} = \frac{-0.5\hat{x} + 1.0\hat{y} - 1.0\hat{z}}{1.5}$. 
We begin our analysis with defining 4 ports at the terminals. First, $4 \times 4$ S-parameter matrix is extracted via eigen analysis for the corresponding ports. Next, ports are terminated with $50\Omega$ and full wave analysis is performed with external field excitation. Subsequently, hybrid S-Parameters are extracted from the full wave analysis with plane wave excitation and $4 \times 4$ S-parameters are updated with the obtained hybrid S-parameters to construct $5 \times 5$ hybrid S-matrix.

Lastly, the corresponding $5 \times 5$ hybrid S-matrix is exported to ADS and circuit analysis is employed for each configuration shown in Fig. ?? To compare the proposed method results with full wave results, we also performed full wave analysis for each configuration and the results for each case is given in Fig. 3.13. In this plot, voltage magnitude and phase are compared at each port and it is obviously seen that proposed method agrees with full wave solution reasonably well. 

![Figure 3.12: Pair of microstrip lines residing on a RT/Duroid 5870 PCB and corresponding port terminations](image)

<table>
<thead>
<tr>
<th>Port #</th>
<th>Configuration: 1</th>
<th>Configuration: 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>P1</td>
<td>$R=50 \Omega$</td>
<td>$R=70 \Omega$</td>
</tr>
<tr>
<td></td>
<td>$C=0.5 \text{ pF}$</td>
<td>$L=1 \text{ nH}$</td>
</tr>
<tr>
<td></td>
<td></td>
<td>$C=2.5 \text{ pF}$</td>
</tr>
<tr>
<td>P2</td>
<td>$R=300 \Omega$</td>
<td>$R=120 \Omega$</td>
</tr>
<tr>
<td></td>
<td>$C=2.0 \text{ pF}$</td>
<td>$L=2 \text{ nH}$</td>
</tr>
<tr>
<td></td>
<td></td>
<td>$C=3.5 \text{ pF}$</td>
</tr>
<tr>
<td>P3</td>
<td>$R=215 \Omega$</td>
<td>$R=270 \Omega$</td>
</tr>
<tr>
<td></td>
<td>$C=1.0 \text{ pF}$</td>
<td>$L=1 \text{ nH}$</td>
</tr>
<tr>
<td></td>
<td></td>
<td>$C=2.0 \text{ pF}$</td>
</tr>
<tr>
<td>P4</td>
<td>$R=75 \Omega$</td>
<td>$R=320 \Omega$</td>
</tr>
<tr>
<td></td>
<td>$C=3.0 \text{ pF}$</td>
<td>$L=2 \text{ nH}$</td>
</tr>
<tr>
<td></td>
<td></td>
<td>$C=2.0 \text{ pF}$</td>
</tr>
</tbody>
</table>
3.2.5 Field Coupling to a Multilayer PCB

We continue our validation with a multilayer PCB subject to a plane wave excitation at 2 GHz (see Fig. 3.14).

Geometry and Excitation Details:

$E_x = 500V/m$ and $E_y = 750V/m$

$\hat{k} = \frac{0.75\hat{x} - 0.5\hat{y} - 1.0\hat{z}}{1.23}$

Bottom substrate $\varepsilon_r = 2.33$

Top substrate $\varepsilon_r = 3.5$

Bottom substrate thickness=0.31mils

Top substrate thickness=0.31mils

Port 1 and port 2 are placed on the bottom substrate while port 3 and 4 are on the top plate. As in the previous case, $4 \times 4$ S-parameter matrix defined for
PCB ports are modified with hybrid S-parameters and the resulting $5 \times 5$ hybrid S-matrix is exported to ADS and circuit simulations are performed for two different configurations given in Fig. 3.14.

![Figure 3.14: Multilayer PCB subject to plane wave illumination and corresponding port terminations](image)

The proposed method is compared with full wave analysis and the results in Fig. 3.15 shows obvious agreement of the proposed method with full wave results.

### 3.3 Field Coupling to a Multilayer PCB inside a Cavity

Next, we study a more realistic structure as in Fig. 3.16 displaying a multilayer PCB inside a cavity subject to a plane wave excitation at 2 GHz.

**Geometry and Excitation Details:**

- $E_x = 1000V/m$ and $E_y = 5000V/m$
- $\hat{k} = \frac{0.5\hat{x} - 1.0\hat{y} - 1.0\hat{z}}{1.5}$
- Bottom substrate $\varepsilon_r = 2.33$
Figure 3.15: Comparison of the proposed hybrid S-matrix method with full wave analysis for the geometry in Fig. 3.14

Top substrate $\varepsilon_r = 3.5$

Bottom substrate thickness=0.31mils

Top substrate thickness=0.31mils

Cavity dimensions 12x10x4cm

Aperture radius 1cm

Port 1 and port 2 are placed on the bottom substrate and port 3 and 4 are on the top plate. As in the previous case, $4 \times 4$ S-parameter matrix defined for PCB ports are modified with hybrid S-parameters and the resulting $5 \times 5$ hybrid S-matrix is exported to ADS and circuit simulations are performed for two different configurations shown in Fig. 3.16.

In Fig. 3.17, comparison of the proposed method results with full wave analysis is given. It is clearly observed that proposed method agrees very well with full wave results even when PCBs are housed inside cavity-like structures.
3.4 Wide Band Field Coupling to a PCB Designed for a Power Amplifier inside Cavity

In this example, we study field coupling to a PCB board designed for RF Power amplifier and located inside a cavity subject to field excitation (see Fig. 3.18). We defined four ports at the input and output of the board in addition to the ports at the gate and drain of the amplifier.

As in the previous cases, we modeled the whole board with $4 \times 4$ S-matrix and conducted hybrid S-parameter analysis to extract hybrid scattering entries. Subsequently, new hybrid $5 \times 5$ S-matrix was exported to ADS and loads connected to the corresponding ports (see Fig. 3.18). PCB inside the cavity was subject to wideband EMI ranging from 2GHz to 3GHz. We compared hybrid S-parameter results with full wave solution (HFSS) and voltage magnitude and phase at ports 1 and 2 in Fig. 3.19.
Figure 3.17: Comparison of the proposed hybrid S-matrix method with full wave analysis for the geometry in Fig. 3.16

Figure 3.18: External field coupling to a PCB inside a cavity for a wide range of frequencies: This study was performed by Zulfiqar Khan
Figure 3.19: Comparison of the proposed hybrid S-matrix method with full wave analysis for the geometry in Fig. 3.18: This validation study was performed by Zulfiqar Khan clearly indicates that hybrid S-parameters agree very well with full wave solution for a wide frequency range.

3.5 Field Coupling to an Inverter inside a Cavity

In this section, we investigate output characteristics of the same TI inverter used in Chapter 2 when it is subject to off-board EMI. The geometry in Fig. 3.20 is used and corresponding $9 \times 9$ broad band hybrid S-parameter matrix is exported to HSPICE. Inverter input, output and DC power input are connected to the ports 4, 5 and 6 respectively.
Figure 3.20: Inverter inside a metallic box for off-board EMI analysis

Figure 3.21: Investigation of off-board EMI effects on a TI inverter (see Fig. 3.20 via hybrid S-Parameter analysis)
In this example, the inverter is operating at 1 GHz with pulse width of 300 ps. External EMI is operating at 2 GHz and the plane wave magnitude of 1414 V/m and 2828 V/m are applied. It has been clearly seen in Fig. 3.21 that external EMI generates deteriorating effects on the output of the inverter and this effect becomes more dominant with increasing plane wave magnitude, thereby, leading to system level upset. This example also implies that even though inverter is shielded, a small aperture cut-out for input and output cables may lead to severe coupling to the device, and eventual logic failure.
CHAPTER 4

HIGH POWER EMI ON RF POWER AMPLIFIER AND DIGITAL MODULATION SCHEMES

Having developed the hybrid S-parameters technique in the previous chapter, we now employ the port modeling in conjunction with hybrid S-parameters to address the impact of IEMI on communication systems and devices.

Communication systems are a fundamental part of many civil and military applications. Therefore, understanding their susceptibility to IEMI plays a critical role in designing and protecting existing and next generation systems. In this chapter, we focus on characterizing system and device level upsets arising in communication electronics when they are subject to on and off – board IEMI. We particularly carry out experimental and theoretical studies for performance analysis of an RF power amplifier and digital modulation schemes when exposed to high power microwave injection or attacks. For the computational analysis, we employ hybrid S-parameter method to address off – board IEMI attacks and port modeling method (see Chapter 2) for on – board EMI analysis with the goal of characterizing and validating the carried out measurements.

Digital modulation systems are prevalent in communication systems due to their higher information capacity, better quality (better noise performance) and higher
data security. As a result of these advantages, there is a continuous transition from analog and frequency modulation systems to digital modulation techniques including QPSK (Quadrature Phase Shift Keying), FSK (Frequency Shift Keying), MSK (Minimum Shift Keying) and QAM (Quadrature Amplitude Modulation) [101]. These modulation techniques represent core component of Digital Video Broadcasting, wireless communications such as GSM and CDMA and many other defense and aerospace applications. Their performance assessment under high IEMI environments, therefore, constitutes a critical issue for the current and next generation designs and deployments.

RF power amplifier is a fundamental component of any communication system and prevalently used to amplify modulated signals. Therefore, its susceptibility against intentional EMI plays an important role in assessment of performance of the communication systems when subject to high power IEMI attacks.

In the subsequent section, we start with describing the impact of IEMI on RF power amplifier device characteristics such as power gain. We discuss experimental observations and compare with our theoretical analysis for on-board EMI analysis. We next extend this analysis to a more complex setup in which the amplifier is placed inside a missile-like body (hollow cylindrical enclosure) and subject to high power single and multi tone intentional EMI. Subsequently, we employ the hybrid S-parameter technique to carry out computational analysis for the aforementioned setup.

In the last section of the chapter, we provide experimental and theoretical IEMI evaluations of digital modulation schemes applied through a power amplifier. For the digital modulation schemes, we particularly consider constant (phase-modulated)
and non-constant envelope (amplitude-modulated) modulation schemes to evaluate their performance (Error Vector Magnitude, Upset or Failure) due to IEMI attacks. We further employed the hybrid S-parameters for off-board EMI analysis of digital modulation systems deployed inside a missile-like body.

4.1 Impact of High Power EMI on Device Characteristics of an RF Power Amplifier

We begin our analysis by investigating the impact of high power IEMI on the characteristics of an RF power amplifier designed in collaboration with the Non-linear RF Laboratory at The Ohio State University.

Input and output matching, drain and gate bias networks for a Free Scale MRF281S class AB amplifier on a RT/Duroid 5880 board was designed in-house (see Fig. 4.1). The design was optimized to operate in GSM and W-CDMA bands and biased with 26V at the drain and 5.4V at the gate. The amplifier yielded about 20dB and 17dB transducer power gains in the GSM and W-CDMA bands, respectively (see Fig. 4.1(b)).

For computational analysis, we carried out wide-band (up to 8.4GHz) port characterization of the RT/Duroid board to generate 15×15 S-parameter matrix at corresponding matching and bias ports via ADS Momentum (2.5D Method of Moments simulation tool). The power amplifier, the matching and bias network components were connected at the corresponding ports and simulated via ADS Harmonic Balance Simulator for the power gain analysis. Comparison of measurement and simulations for the transducer power gain with an input power of 5dBm is illustrated in Fig. 4.1(b).

2The author is grateful to Dr. Suk Keun Myoung, Saek Joo Doo and Prof. Patrick Roblin for their help and guidance during the course of studies presented in this chapter.
Figure 4.1: (a) FreeScale MRF281S class AB RF power amplifier on RT Duroid 5880 Board-(b) Amplifier transducer power gain ($P_{\text{avg}} = 5\text{dBm}$, $V_{DS} = 26V$ and $V_{GS} = 5.4V$)-(c) RF power amplifier design diagram
As shown, measurements and simulation agree well, validating the accuracy of the theoretical modeling of the amplifier network and the circuit board (see Fig. 4.2).

4.1.1 Experimental and Theoretical Analysis of On-Board EMI on RF Amplifier Gain

Having shown that our port modeling can capture EM effects on the PCB, we next proceed to investigate on-board EM effects on the amplifier gain via experimental and theoretical approaches.
We first took a series of measurements in Non-linear RF Laboratory to characterize the amplifier’s gain with and without IEMI. The measurement setup, displayed in Fig. 4.3, was comprised of the Agilent ESC4438C for signal generation, Anritsu MG3692A CW Generator for EMI signal, LSNA and MATLAB post-processing for capturing and analyzing the measurement data.

Our first set of measurements aiming to identify the amplifier 1dB compression point (ie. the output power level when the amplifier begins to operate in the non-linear mode) yielded 35dBm at the output with the input power of 20dBm at 0.940 GHz (see Fig. 4.4(a)). Stated differently, the amplifier starts to operate in the non-linear mode when the input power is greater than 20dBm at 0.940GHz.

We next measured the degradation in the output power and gain of the amplifier when high power EMI is injected into the gate. Specifically, we injected EMI signal at power levels of 8dBm, 18dBm and 23dBm all at 0.949GHz at the input terminal. As shown in Fig. 4.4(a)-(b), the amplifier output power and gain significantly dropped by about 8dB when the device is subject to 23dBm of EMI power. This phenomenon is also referred to as desensitization that a very large interferer can lead to vanishingly small gain for a weak input signal. We further observed the same conclusions with port modeling approach run via ADS Harmonic Balance Simulator (see Fig. 4.2).

Substantial degradation in the device gain can be attributed to the power of fundamental harmonic distributed over intermodulation products since the device is forced into non-linear mode due to high power EMI coupled to the amplifier. High power EMI at the device terminals also heats up the device very quickly leading to further performance degradation. More specifically, we observed that when the EMI power was 23dBm, DC current at the drain was 0.5mA, viz almost twice the
Figure 4.3: Setup for the measurement of RF amplifier device characteristics subject to high power EMI

DC current in the absence of EMI (0.27mA). Due to the large DC drain power, the amplifier’s efficiency decreased significantly and the device heated up fast as a result of very high DC power at the terminals.

To sum up our observations in this section, we conclude that even if the amplifier is operating in the linear region for its normal operation, its performance can be severely compromised with high power EMI attack. And, this may eventually lead to permanent device upset depending on the duration of the EMI attack.
Figure 4.4: RF amplifier subject to high power EMI: (a) Output power vs. input power (b) Gain vs. input power

4.1.2 Off – Board High Power EMI on the Characteristics of an Enclosed RF Amplifier

We demonstrated in the previous section that very high EMI power is needed at the amplifier terminals to force the device into the non-linear region causing significant decrease in the power gain. In this section, we shield the amplifier and the PCB by placing them inside a missile-like body to investigate what power is required in a more realistic platform to cause noticeable degradation in the power gain (see Fig. 4.5) when the device is subject to intentional EMI from a certain distance. To do so, we designed a metallic enclosure with a small apertures on the shielding box for air ventilation purpose. The metallic enclosure was placed inside a missile-like body and exposed to high power EMI from a source placed at 15m distance (see Fig. 4.5). The
Figure 4.5: (a) RF Power amplifier shielded with a metallic box inside a missile-like body with wings attached - (b) Missile-like body and EMI source configuration

EMI source was at 1.0GHz and had an antenna gain of 20dBi in the direction towards the subject.

For computational modeling, we employed the hybrid S-parameter method to incorporate plane wave coupling with on-board S-parameters. In particular, we adopted the simulation approach in Fig. 4.6. EMCAR was used to compute the transfer function for the missile-like body. This analysis assumed that the interaction between the missile-like body and the metallic box was at minimal level such that it would not
Figure 4.6: Simulation approach for off-board EMI analysis of power amplifier inside the missile-like structure

significantly alter the conclusions we could draw from this analysis. Considering that the height of the metallic box is very small compared to the missile-like body, such an assumption is acceptable. In other words, our analysis in this section accounts for only first order coupling from the EMI source to the metallic enclosure inside the missile-like structure.

In evaluating the plane wave magnitude impinging on the missile-like body, we assumed free space radiation ignoring interactions between the EMI source and the missile-like body. Thus, the magnitude of the plane wave illuminating the body was computed with the following formula

\[
\frac{P_t G_t}{4\pi d^2} = \frac{|E|^2}{2Z_0}
\]  

(4.1)

where \(P_t\) and \(G_t\) refer to the EMI source power and antenna gain, respectively.

Once we compute the transfer function between the missile-like body and the EMI source via Method of Moments (EMCAR), we next employ hybrid S-parameter method for the external field coupling to the PCB inside the metallic box. HFSS was used to extract the hybrid S-matrix entries. Having computed the missile transfer
function and the hybrid S-parameters, we subsequently combined all the parameters in a single hybrid S-matrix and exported to ADS for harmonic balance analysis (see Fig. 4.6).

Referring to Fig. 4.7(a), we display the amplifier transducer power gain for the frequency range of 0.01-3GHz when exposed to a single tone EMI at 1GHz with various power levels from a source at 15m distance.

Fig. 4.7(a) shows that the EMI power of 50MW decreases the amplifier gain by 5dB for a wide frequency range. Such high EMI power corresponds to peak electric field magnitude of 36.5kV/m at 15m from the source in the absence of the missile-like structure. Degradation in the gain even increases further by 8dB when EMI power is raised to 100MW corresponding to 51.6kV/m peak field value. Despite such high power levels required to have adverse effect on the amplifier performance, we must
state that similar realistic power sources have been already used in practice to study susceptibility of electronic systems [2].

Our next study investigates whether multi-tone EMI signals with the same total power level (as in the single tone) lead to the same impact on the amplifier gain. To do so, we use five tone modulated EMI signal with each tone separated by 10MHz and centered around 1GHz. The simulation results in Fig. 4.7(b) indicates that multi-tone EMI has less impact on the gain of the amplifier than single tone EMI does even if the total average power radiated by the EMI source is the same for the both. In other words, since the number of intermodulation products increased with the coupling of multi tone EMI, the power is distributed over more intermodulation products. As a result, it has less impact on the fundamental operational harmonic as compared to the single tone EMI.

4.2 Performance of Digital Modulation Schemes under High Power IEMI

Our analysis above is based on the evaluation of device level upsets when the amplifier is exposed to high power IEMI. In this section, we evaluate the performance of digital modulation schemes such as Quadrature Phase Shift Keying (QPSK), Minimum Shift Keying (MSK), Quadrature Amplitude Modulation (QAM) and binary modulation schemes such as Binary Phase Shift Keying (BPSK) by computing the Error Vector Magnitude (EVM) when they are processed through RF amplifier subject to IEMI (see Appendices A and B for overview of digital modulation schemes and Error Vector Magnitude, respectively). Both in-band and out-of-band EMI signals are used to excite the system for the vulnerability evaluation of the aforementioned modulation schemes. We aim to evaluate the deterioration of the modulated signals
and possible upsets on the electronic modules that process these schemes. To do so, an RF-power amplifier was employed prior to the transmission stage to assess device effects on the modulated signal by measuring the EVM. Our objectives in this section is to identify

- Which modulation schemes are more susceptible to IEMI
- Which modulation features play a critical role in the system upset
- How much power is needed to fail a typical communication system enclosed by complex realistic bodies such as a missile-like structures

We employed the same power amplifier shown in Fig. 4.1 and carried out the measurements in Non-linear RF laboratory for on – board EMI analysis of the digital modulation schemes. The measurements are also validated with theoretical analysis using the ADS Ptolemy cosimulation package with envelope (mixed time-frequency) analysis. We also carried out off – board IEMI analysis of the digital modulation system enclosed by the missile-like structure. This analysis was done via combined ADS Ptolemy cosimulation and the hybrid S-parameter method.

Below, we first present measurement setup and the theoretical approach used for the validation and off – board EMI analysis. We next consider on – board EMI measurements and analysis for each modulation scheme. This is followed by off – board EMI analysis of digital modulation systems enclosed by a missile-like structure.

4.2.1 Measurement Setup for the Digital Modulation Schemes Subject to High Power EMI

The measurement setup for the on – board EMI evaluation of digital modulation schemes, shown in Fig. 4.8, was comprised of an Agilent 4438C for digitally modulated
signal generation, Anritsu CW signal generator for the EMI signal and the LSNA unit with MATLAB for the data capturing and post-processing.

We had a random 64 bit digital data sequence hard-coded in the Agilent Vector Signal Generator with an RF carrier frequency of 0.94GHz at various signal powers (input to the RF amplifier). We note that our study initially aimed to evaluate digital communication systems carried in both GSM and W-CDMA bands. However the RF carrier frequency was specifically chosen in the GSM band since the equipment available to us for the EMI signal amplification was limited to operate at a
maximum of 1GHz. Even though EMI power required to upset systems operating in W-CDMA band might be different than what is presented in the subsequent sections, we must note that the overall conclusion in terms of EMI source characteristics will not be impacted by performing the study only in GSM band.

4.2.2 Simulation Setup

The ADS ptolemy package was used for computational analysis. It is comprised of modulation and demodulation tools that can cosimulate RF digital systems concurrently. Referring to Fig. 4.9, we show ADS circuit model for each component used in the measurement. The first section in the schematic is comprised of modulation components which take the digital data and split into in-phase and quadrature phase components. The components in this section are specifically created for each modulation scheme and developed for custom use to match the modulation algorithms used by the ESG4438C. The second stage in the schematic is comprised of amplifier with its matching and bias network models as given in Fig. 4.2 and a multi-tone signal generator for the EMI. The last stage represents data capturing and demodulation for post processing. Finally, the EVM evaluation was done in MATLAB by processing the digital data captured at the last phase of the schematic in Fig. 4.9.

We again modeled the PCB with a broadband S-parameter network and circuit elements were connected at the corresponding ports.

For off-board EMI analysis, EMCAR was used to carry out full-wave analysis of the missile-like body. The generated fields at the PCB locations were then used in connection with the hybrid S-parameter matrix enclosing the amplifier (see Fig. 4.6).
Figure 4.9: Simulation setup for on-board EMI analysis of digital modulation schemes
Using the above setup, we next proceed with the experimental and theoretical validation for each modulation scheme when exposed to on&off – board IEMI.

### 4.2.3 Performance of the Quadrature Phase Shift Keying in Presence of IEMI

The Quadrature Phase Shift Keying (QPSK) is a constant amplitude modulation scheme employed primarily for Digital Video Broadcasting-Satellite (DVB-S), Code Division Multiple Access (CDMA) and Wireless Local Loop applications. In accordance with the QPSK scheme, an input pair of bit stream is split into two separate bits and each is modulated with orthogonal cosine and sine functions to generate in-phase (I) and quadrature phase (Q) components, respectively (see Fig. 4.10 (a)-(b)). Due to the orthogonality of the cosine and sine functions, I/Q components are easily demodulated by the same orthogonal functions at the receiving end. Constellation diagram in Fig. 4.10(b) is used at the receiving end for data regeneration depending on where the received I/Q signals falls on the diagram. Deviation and transition of the signals on the I/Q map can be used to evaluate the overall performance of the modulation scheme (see Appendix A for more detailed overview of QPSK).
To investigate the performance of our computational model for the RF amplifier and the QPSK characterization, we carried out measurements in the absence of EMI when the RF power amplifier operates in the linear and non-linear modes. In the measurement setup, the Vector Signal Generator was programmed to generate 9dBm signal power (input power to the amplifier) for 64-bit randomly generated digital bit stream within 20MHz bandwidth and at a carrier frequency of 0.940GHz. We, subsequently, captured the amplifier output via LSNA and post-processed for the demodulation. The same analysis was also carried out via ADS using the model described in Fig. 4.9. Comparing the measurements for the input/output transition diagram in Fig. 4.11(a) with the simulations in Fig. 4.11(b), we can readily see that our simulation results match very well with the measurements thus validating our computational model.

We continue our analysis by increasing the input power so that the RF amplifier is driven into the non-linear region leading to higher harmonics to arise. Noting that the RF amplifier has 1dB compression point at the input power of 20dBm, we used 22dBm output power for 64-bit random bit sequence each pulse with 0.5\(\mu\)s width. The corresponding measurement and simulation results are given in Fig. 4.12(a)-(b) showing that the device non-linearity does not significantly affect the transition diagram. Even though its non-linearity introduces additional intermodulation products within the bandwidth, their power levels are so small that the output data stream is affected negligibly. This observation hints that forcing the device into non-linear region is not a very effective approach to impact the performance. This is further investigated in the later sections below.
Figure 4.11: (a) Measurements displaying the transition diagram for QPSK (RF amplifier is in linear region); RF amplifier input @9dBm - (b) Simulation results for the same setup
Figure 4.12: (a) Measurements displaying the transition diagram for QPSK (RF amplifier is in non-linear region); RF amplifier input @22dBm - (b) Simulation results for the same setup.
We next exposed the amplifier to on-board EMI at the gate terminal. We used the same data sequence in the previous analysis and injected EMI power of -5dBm and 0dBm at 0.945GHz at the amplifier gate, respectively. A notable impact in the output transition diagram is observed with the presence of -5dBm EMI power (see Fig. 4.13(a)). Even though significant deviation occurred from the ideal transition diagram when the EMI was introduced, we observe that the modulation works properly since deviations are still within decision boundaries. The deviation is further exacerbated when EMI power is increased to 0dBm (see Fig. 4.14(a)) implying high likelihood of the improper system functioning when the output is demodulated. These observations are also verified with the simulations in Figures 4.13(b) and 4.14(b).

A more standard approach to characterize the EMI effects on the QPSK modulation is to evaluate EVM after the EMI is injected. The measurement and simulations yielded RMS EVM of 0.33 and 0.27, respectively for -5dBm EMI power at the gate (see Fig. 4.15(a)-(b)). RMS EVM even further increased to 0.58 and 0.50 (measurement and simulations, respectively) for EMI power of 0dBm as shown in Fig. 4.14(a)-(b). Noting that 3G standards require RMS EVM of 0.17 or less for QPSK, we can readily conclude that in-band EMI at practically achievable power levels cause significant degradation in the performance of communication systems employing QPSK [102].

In addition to the in-band EMI, we also conducted measurements to characterize out-of-band EMI on the performance of the digital modulation schemes. To do so, we coupled very large EMI power to the input terminals (about 20dBm at 1.885GHz-second harmonic of carrier frequency). In spite of this higher power, we did not observe any difference at the output although the device was forced to operate in nonlinear mode. In other words, our observations for the output I/Q transition
Figure 4.13: (a) Measurements displaying the transition diagram for QPSK (EMI @-5dBm and 0.945GHz); RF amplifier input @5dBm - (b) Simulation results for the same setup.
Figure 4.14: (a) Measurements displaying the transition diagram for QPSK (EMI @0dBm and 0.945GHz); RF amplifier input @5dBm - (b) Simulation results for the same setup
Figure 4.15: (a) Measured EVM for the QPSK modulation scheme with EMI @-5dBm (0.945GHz) (b) Simulation results for the same setup

Figure 4.16: (a) Measured EVM for the QPSK modulation scheme with EMI @0dBm (0.945GHz) (b) Simulation results for the same setup
off-board IEMI analysis of QPSK via hybrid S-parameters

We have concluded in the previous section that on-board EMI power of 0dBm at the amplifier terminals would suffice to disrupt the operation. To associate such power levels with practical systems, we placed the amplifier inside the missile-like body and exposed to high power EMI from a 15m distance (see Fig. 4.17). We again shielded the amplifier with a metallic enclosing box as done previously and exposed to single tone EMI source operating at 0.945GHz with an antenna gain of 20dBi.

We used the same 64 bit data sequence and modulated at 0.940GHz with an output power of 5dBm. For the analysis, we again employed the hybrid S-parameter technique to account for external field coupling onto the amplifier terminals (see Fig. 4.6).

Referring to Fig. 4.18(a)-(b), we observe that the RMS EVM with an average power of 100kW and 1000kW generates similar distortions to those already achieved
Figure 4.18: (a) QPSK transition diagram and EVM for the setup in Fig. 4.17 (a). EMI power of 100kW @ 0.945GHz - (b) EMI power of 1000kW @ 0.945GHz

with 0dBm in-band EMI coupled to the amplifier gate. Comparing the impact 1000kW of in-band external EMI power generated with 100MW required to generate about 8dB degradation in gain of the amplifier, we conclude that failing digital modulation schemes run through an RF amplifier requires less resources than upsetting the amplifier to cause similar impact.

We next continue with the same analysis for Minimum Shift Keying modulation scheme.
4.2.4 Performance of the Minimum Phase Shift Keying in Presence of IEMI

Minimum Shift Keying (MSK) scheme is proposed to overcome abrupt phase changes in the QPSK that leads to design challenges. Similar to QPSK, it is a constant envelope modulation scheme (phase modulated) and has in-phase and quadrature components that are shaped by another set of cosine and sine functions to smooth transitions in the I/Q map (see Fig. 4.19). MSK is primarily used in the popular Global System for Mobile Communications (GSM) where Gaussian waveform is employed instead of cosine and sine functions to achieve smoother transitions, thus better spectrum. It is also referred as Gaussian Minimum Shift Keying (GMSK). The reader is referred to Appendix A for more details about the MSK.
On - Board Experimental and Theoretical IEMI Analysis of MSK

We carry out the same above on&off – board EMI analysis for the MSK modulation scheme running through the RF amplifier.

We again injected amplifier gate in-band EMI of 5dBm and 0dBm and captured in-phase and quadrature components for the EVM analysis using the setup in Fig. 4.8. Referring to Fig. 4.20 displaying output I/Q transition diagram and Fig. 4.21 for EVM over time when the system is exposed to -5dBm of EMI power at 0.945GHz, we observe RMS EVM of 0.25 and 0.23 according to experimental and computational analysis, respectively. Although the EMI caused noticeable disturbance in the output transition diagram, the system is still operational since the output signal remains within the decision boundaries.

We next increased EMI power to 0dBm and observed that higher EVM of 0.4 (see Figures 4.22 and 4.23). Hence, MSK has better performance than QPSK under the same EMI power. Even though this is not very clear in the transition diagram whether the system would fail under this scenario given the very high EVM obtained, we expect to see large Bit Error Rate when the EMI power of 0dBm is applied to the input terminal of the amplifier.

Off - Board IEMI Analysis of MSK via Hybrid S-Parameters

Our focus in this section is to investigate how much external EMI power is needed to have noticeably adverse impact on the MSK modulation for the setup given in Fig. 4.17.

Referring to the Fig. 4.24(a)-(b) displaying the output transition diagram and the corresponding RMS EVM for EMI sources having average power of 100kW and
Figure 4.20: (a) Measurements displaying the transition diagram for MSK (EMI @-5dBm and 0.945GHz); RF amplifier input @5dBm - (b) Simulation results for the same setup
1000kW with 20dBi antenna gain, respectively, we observe that the latter was able to generate as comparable level of upset as on-board EMI of 0dBm does. This analysis also emphasizes the observation we had for the QPSK that upsetting digital systems processed through RF amplifier requires less resources than causing device level upset to impact the system performance unfavorably.

4.2.5 Performance of the Binary Phase Shift Keying in Presence of IEMI

Binary Phase Shift Keying (BPSK) is a constant envelope modulation technique and one of the most robust modulation schemes since the decision boundary is further separated in the I/Q map (see Fig. 4.25). It is mainly employed for applications in Deep-Space Telemetry, Wi-Fi and cable modems. A more detailed overview is given in Appendix A.
Figure 4.22: (a) Measurements displaying the transition diagram for QPSK (EMI @ 0dBm and 0.945GHz); RF amplifier input @5dBm - (b) Simulation results for the same setup
Figure 4.23: (a) Measured EVM for the MSK modulation scheme with EMI @0dBm (0.945GHz) (b) Simulation results for the same setup

On-board Experimental and Theoretical IEMI Analysis of BPSK

We carried out on-board EMI measurements in which EMI power of -5dBm and 0dBm at in-band frequency of 0.945GHz is coupled to the input terminals of the amplifier as previously described in Fig. 4.8.

Referring to the I/Q transition diagram in Fig. 4.26 and EVM versus time in Fig. 4.27 for EMI power of -5dBm, we note RMS EVM of 0.22 and 0.18 from experimental and computational analysis, respectively.

We next increased EMI power to 0dBm and observed relatively higher RMS EVM of 0.38 which is comparable with MSK modulation scheme when it is subject to the same EMI power (see Figures 4.28 and 4.29). However, it is clear in the transition diagram that the system would still be operational, despite the very high EVM obtained.
Figure 4.24: (a) MSK transition diagram and EVM for the setup in Fig. 4.17 (a) EMI power of 100kW @0.945GHz -(b) EMI power of 1000kW @0.945GHz

Figure 4.25: (a) BPSK modulation- (b) BPSK constellation diagram
Figure 4.26: (a) Measurements displaying the transition diagram for BPSK (EMI @-5dBm and 0.945GHz); RF amplifier input @5dBm - (b) Simulation results for the same setup
Off - Board IEMI Analysis of BPSK via Hybrid S-Parameters

Having analyzed performance of BPSK under high power on – board EMI, we now employ the hybrid S-parameter method to find out how this modulation scheme would respond to an external EMI when the amplifier is protected by a metallic box enclosed by the missile-like body (see Fig. 4.17). The same EMI source as in previous modulation schemes was used with single tone power and antenna gain of 20dBi at 0.945GHz. We also used the same 64 bit data sequence carried at 0.940GHz with output power of 5dBm.

RMS EVM for EMI sources with 100kW and 1000kw power is 0.14 and 0.45 respectively as given in Fig. 4.30(a)-(b). When this is compared with MSK and QPSK for the same external EMI power, we readily see that BPSK has better performance.
Figure 4.28: (a) Measurements Displaying the transition diagram for BPSK (EMI @0dBm and 0.945GHz); RF amplifier input @5dBm - (b) Simulation results for the same setup.
Figure 4.29: (a) Measured EVM for the BPSK modulation scheme with EMI @0dBm (0.945GHz) (b) Simulation results for the same setup

Figure 4.30: (a) BPSK transition diagram and EVM for the setup in Fig. 4.17 (a) EMI power of 100kW @0.945GHz -(b) EMI power of 1000kW @0.945GHz

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4.2.6 Performance of Quadrature Amplitude Modulation Method in Presence of IEMI

In this section, we focus on non-constant envelope (amplitude modulation) scheme Quadrature Amplitude Modulation Method 16 (QAM16). Similar to QPSK, QAM16 has in-phase and quadrature components that are modulated by cosine and sine functions. The distance between constellations points are so small in QAM16 that it is more susceptible to noise (see Fig. 4.31). Due to its higher data capacity, QAM16 has many applications including Microwave Digital Radio, Wi-Fi, Digital Video Broadcasting-Cable (DVB-C) and Digital Video Broadcasting Terrestrial (DVB-T). A more detailed explanation of the QAM16 is given in Appendix A.
On - Board Experimental and Theoretical IEMI Analysis of QAM16

Similar to the previous analysis, we used 64 bit random data sequence with each bit having 0.5µs pulse width. We injected EMI powers of -5dBm and 0dBm at 0.945GHz to the gate terminal of the amplifier (see Fig. 4.8).

Referring to Figures 4.32 and 4.33 displaying output I/Q transition diagram for -5dBm and 0dBm, respectively, we observe significant deviations in the I/Q transition map. Due to the small decision boundaries, we can easily state that QAM16 is more susceptible to the EMI than any other modulation schemes discussed here. This is further substantiated with the RMS EVM we observed in Figures 4.34 and 4.35.

Considering the Fig. 4.34 for RMS EVM over time when the system is exposed to -5dBm of EMI power at 0.945GHz, we observe RMS EVM of 0.34 and 0.32 according to experimental and theoretical analysis. It further increased to 0.59 and 0.50, according to measurements and computational analysis, respectively (see Figures 4.35 for EMI power of 0dBm. Noting that 3G Conformance standards suggest RMS EVM of .125 or less for QAM16 modulation scheme, we conclude that this modulation system is the most susceptible modulation scheme amongst the ones we analyzed in this chapter.

Off - Board IEMI Analysis of QAM16 via Hybrid S-Parameters

We next proceed with off – board EMI analysis of QAM16 using the setup in Fig. 4.17).

Referring to the simulation results in Fig. 4.36(a)-(b) for EMI powers of 100kW and 1000kW , respectively, we observe that the system was upset significantly even at 100kW. Even though we observe a low RMS EVM of 0.33 for 100kW of EMI power, the system is unable to perform properly since distorted transition diagram
Figure 4.32: (a) Measurements displaying the transition diagram for QAM16 (EMI @-5dBm and 0.945GHz); RF amplifier input @5dBm - (b) Simulation results for the same setup.
Figure 4.33: (a) Measurements displaying the transition diagram for QAM16 (EMI @0dBm and 0.945GHz); RF amplifier input @5dBm - (b) Simulation results for the same setup.
Figure 4.34: (a) Measured EVM for the QAM16 modulation scheme with EMI @-5dBm (0.945GHz) (b) Simulation results for the same setup.

Figure 4.35: (a) Measured EVM for the QAM16 modulation scheme with EMI @0dBm (0.945GHz) (b) Simulation results for the same setup.
runs over the decision boundaries. Therefore, it causes significantly high bit error rate. Compared to the other modulation schemes we investigated above, we note that QAM16 is much more susceptible to EMI thus making it an easy target for low power IEMI sources.
4.3 Performance Comparison of Constant and Nonconstant Envelope Digital Modulation Schemes Subject to IEMI

In this section, we discuss performance of constant envelope modulation schemes (QPSK, MSK, BPSK) with non-constant envelope modulation methods (QAM16) exposed to on&off – board EMI.

Referring to the Fig. 4.37 which is comparing RMS EVM for all the modulation schemes when EMI power of -5dBm and 0dBm is injected to the gate of the amplifier, we notice that constant envelope modulation schemes perform better than the non-constant envelope modulation scheme(QAM16). The same observation is further substantiated when external EMI is coupled to the amplifier inside missile-like body (see Fig. 4.38). Among constant envelope modulation schemes, BPSK has better performance than MSK and QPSK. Even though MSK is slightly modified version of QPSK, we notice MSK has significantly higher performance than QPSK when exposed tp on – board EMI. However, the performance difference between MSK and QPSK is not noticeable anymore in presence of external EMI. This can be attributed to that external EMI also couples to the drain and the output of the amplifier causing significant degradation in the performance of the MSK modulation scheme.

The aforementioned performance difference among the modulation schemes can also be explained with their inherent noise performances. Depending on the bandwidth efficiency of each modulation scheme, their susceptibility to the noise largely varies. Referring to the Table 4.1 displaying bandwidth efficiency for each modulation scheme, we can clearly see that QAM16 is the most efficient modulation scheme. However, it has the highest susceptibility to the EMI as we observed above.
<table>
<thead>
<tr>
<th>Modulation Scheme</th>
<th>Efficiency</th>
</tr>
</thead>
<tbody>
<tr>
<td>QPSK</td>
<td>2bit/seconds/Hertz</td>
</tr>
<tr>
<td>MSK</td>
<td>1bit/seconds/Hertz</td>
</tr>
<tr>
<td>BPSK</td>
<td>1bit/seconds/Hertz</td>
</tr>
<tr>
<td>QAM16</td>
<td>4bit/seconds/Hertz</td>
</tr>
</tbody>
</table>

Table 4.1: Bandwidth Efficiency Comparison of Digital Modulation Schemes

In other words, our observations are similar to the performance these modulation schemes exhibit when they are transmitted through a white gaussian noise channel. However, we must note that such conclusion may not always hold when they are subject to off-board EMI because coupling to the device appears in all directions thus effecting output performance differently.

Figure 4.37: Comparison of EVM for digital modulation schemes subject to on-board EMI
Figure 4.38: Comparison of EVM for digital modulation schemes subject to off-board EMI
CHAPTER 5

IMPACT OF HIGH POWER EMI ON AN AUTOMOBILE TIMER

In the previous chapter, we demonstrated that the performance of communication systems can be readily degraded at practical EMI power levels. In this chapter, we focus on impact of EMI signal time characteristics on the performance of electronic circuits. We particularly consider digital circuits since positive and negative edge triggered clocks are prevalently used in many digital systems. To associate our findings with a realistic system, we further analyze a digital timer inside an automobile subject to a train of high power Gaussian pulses.

We again employ the hybrid S-parameter method to integrate the EMI induced at the circuit ports with ADS for the digital circuits analysis. For our evaluation of time characteristics, we first varied the Gaussian pulse width and period of a Gaussian pulse train (see Fig. 5.4). We subsequently varied the scan angle to find the best impinging direction for maximizing the EMI effects when the timer resides in an automobile.
5.1 Timer Inside an Automobile Subject to High Power EMI

Consider the setup in Fig. 5.1 displaying a timer mounted on a RT/Duroid 5880 board enclosed with a metallic box behind the engine compartment of an automobile. In our setup, the entire automobile was illuminated with a train of high power Gaussian pulses from the driver’s side ($\phi = 90^\circ$). The impinging wave was polarized with $H_z \neq 0$ and $E_z = 0$.

We employed negative edge triggered timer designed with AMI 0.5$\mu$ MOSIS line (see Fig. 5.2). Since the timer was designed in-house, we had the advantage of using an accurate SPICE model to integrate it with our computational EM and circuit solvers explained in the subsequent section.

$^3$Timer was designed in collaboration with Kyechong Kim and Agis Iliadis of University of Maryland [103]
Fig. 5.3 displays the timer mounted on an RT/Duroid 5880 board with a single clock input (QCLK) and a 4 output ($QT_1 \ldots QT_4$) microstrip lines inside a box with ventilator apertures. The timer has actually three inputs (VCC, Enable and Clock). However, to keep our analysis simple (without compromising accuracy), we did not include VCC and Enable in the study since we do not expect them to play critical role given the large bias voltage (5V).

### 5.1.1 Computational EM and Circuit Tools

In our computational analysis, we used numerical EM and circuit tools to model each component in the problem setup. Specifically, we employed EMCAR (validated with measurements, see [6]) to model the automobile subject to high power Gaussian
1. Narrow or wide Gaussian pulses more effective?

Given the same power within a pulse,
Find out impact of pulse width on performance of digital systems

2. High or low pulse repetition rate?

Given the same power within a frame,
Find out impact of pulse repetition rate on performance of digital systems
Figure 5.5: ADS modeling of the automobile and timer via hybrid S-parameters pulse train. The timer board and the enclosing box were analyzed using HFSS (Ansoft) to extract S-parameters for the board and generate hybrid S-parameters to integrate with ADS. In the final step, the entire composite model (automobile+ timer box+timer board) was combined with ADS platform through hybrid S-parameters to perform time-domain analysis with the timer SPICE model (see Fig. 5.5). Among the numerical tools available to us, EMCAR and HFSS are full-wave frequency domain EM solvers. Whereas, ADS is both time and frequency domain circuit analysis tool.

5.1.2 Analysis

Due to the high complexity associated with the problem at hand, we decomposed our analysis into three major components (see Fig. 5.6). However, since the coupling from the timer box to automobile is small as compared to coupling from automobile to timer, we neglect the interactions from the timer box to the automobile. This
Figure 5.6: Analysis method for EM coupling to the timer inside an automobile

is a reasonable assumption because the timer box size is very small as compared to overall size of the automobile. That is, the total field solution is mainly dominated by scattering from the automobile body.

Our theoretical approach is outlined in Fig. 5.6 displaying the train of Gaussian pulses (carrying 99% energy of an ideal Gaussian pulse) represented in frequency domain through its Fourier transform. At each frequency point, EMCAR simulations were performed with the automobile and field values in the presence of the automobile were excitation to the timer board inside the metallic box.

To integrate timer PCB board and the enclosing box with the timer SPICE model, we carried out port analysis of the timer board in presence of the enclosing box to extract the S-parameters using HFSS. We defined 10 ports at the terminals of each signal line (2 port at the terminals of clock bus, 8 port at the terminals of output
5.2 Results

In our analysis, we placed the EMI source at 15 meters from the driver’s side of the automobile (see Fig. 5.7). EMI source operated at the carrier frequency of 1GHz and had an antenna gain of 20 dBi.

5.2.1 Timer Performance Versus Incident Pulse Width

We next proceeded to investigate the role of EMI pulse width in disrupting the timer operation. To do so, we illuminated the automobile with a train of Gaussian pulses (period of $2\mu s$) at different pulse widths modulated at 1GHz.
Our preliminary studies suggested that the EMI source with 133W of average power and 20dBi of antenna gain at 15m distance would generate sufficient interference to upset the timer operation. In evaluation of plane wave magnitude impinging on the automobile, we assumed radiation in free space ignoring the interaction between the EMI source and the automobile. So, magnitude of the plane wave illuminating the car was computed via (4.1).

We examined three different scenarios by varying the pulse width and keeping the same source power. Below we present each of the scenarios and the corresponding results.

**Scenario 1: Pulse Width\((\sigma)\) is 0.15\(\mu s\)**

Referring to Fig. 5.8(a)-(b), we show the incident Gaussian pulse of width 0.15\(\mu s\) and the timer response at all outputs. The EMI source radiates an average power of 133W and its antenna has 20dBi gain. Such power corresponds to peak field value of 162 kV/m at a 15 m distance from the source in the absence of the vehicle as calculated from (4.1). Even though the timer is subject to a large EMI, we observe that it is still operational and has only slight disruptions at outputs OT2 and OT3 coinciding with the time the Gaussian pulse peaks.

**Scenario 2: Pulse Width\((\sigma)\) is 0.30\(\mu s\)**

We next double the Gaussian pulse width without changing the EMI source power and the antenna gain. Thus, the peak field value at 15m from the antenna becomes 115 kV/m. Referring to Fig. 5.9(a)-(b), we still observe similar effects at the timer output (as in scenario 1) even though the timer is still functioning properly.
Figure 5.8: (a) Incident Gaussian pulse (period of $2\mu s$ and width of $0.15\mu s$) at 15m distance from the EMI source (average power of the source is 133W and antenna gain is 20dBi)-(b) Timer response at the output ports

Figure 5.9: (a) Incident Gaussian pulse (period of $2\mu s$ and width of $0.30\mu s$) at 15m distance from the EMI source (average power of the source is 133W and antenna gain is 20dBi)-(b) Timer response at the output ports
Figure 5.10: (a) Incident Gaussian pulse (period of 2µs and width of 0.45µs) at 15m distance from the EMI source (average power of the source is 133W and antenna gain is 20dBi)-(b) Timer response at the output ports

**Scenario 3: Pulse Width (σ) is 0.45µs**

The peak field value at 15m from the antenna is now down to 94 kV/m. However, from Fig. 5.10(a)-(b), we observe that the timer is no longer operational even though the peak interfering signal magnitude was significantly lower than the previous two scenarios. This clearly hints that pulse duration is critical and may be more important than instantaneous peak field magnitude in disrupting the timer. This observation can be attributed to that broad periodic pulses lead to higher amplitude harmonics since they are more narrow-band (harmonics with higher power). In other words, few harmonics each with higher power is more destructive than more harmonics each with lower power. Wide pulses also interfere with the clock signal for a longer duration that disrupts the timing characteristics when the device is edge triggered.
5.2.2 Timer Performance Versus Incident Pulse Repetition

We continue our studies by investigating the role of Gaussian pulse period in the timer disruption. As a part of this study, we illuminate the automobile with a train of Gaussian pulses at different pulse periods and widths carried at 1GHz. However, each frame carries the same power for comparison purposes and has a duration of 4µs. The EMI source delivers 133W and the antenna gain of 20dBi placed at 15m from the vehicle. Again, we assume free space radiation; ignoring the interaction between the EMI source and the automobile. We proceed to consider pulse period varying with the total EMI frame power remaining the same. In this context, a “frame” corresponds to a time duration which contains a set of pulses.

Scenario 1: Pulse Period \(T\) is 1µs

In this scenario, we have a gaussian pulse train periodic with 1µs and frame duration of 4µs. Each pulse has a 0.1875µs width.

EMI source radiates 133W of average power and has 20dBi of antenna gain. Since there are four pulses per frame, each pulse carries 33.25W average power with peak field of 73 kV/m at 15m distance from the source (in the absence of the vehicle). Referring to Fig. 5.11(a)-(b), we observe that the timer is operating with some small noise at the output.

Scenario 2: Pulse Period \(T\) is 2µs

Again, the EMI source radiates the same power as before and has the same antenna gain of 20dBi that corresponds to peak field value of 73 kV/m at 15m from the source (in the absence of the vehicle). Since the frame power is 133W, each pulse carries
Figure 5.11: (a) Incident Gaussian pulse (period of 1\(\mu s\) and width of 0.1875\(\mu s\)) at 15m distance from the EMI source (average power is 133W and antenna gain is 20dBi)-(b) Timer response at the output ports

66.5W average power. Fig. 5.12(a)-(b) clearly shows that the timer is still operational with the EM effects being minimal even though we doubled the repetition rate.

**Scenario 3: Pulse Period**\( (T) \) **is** 4\(\mu s\)

We now have a repetition rate of 4\(\mu s\) corresponding to a single pulse per frame with 0.75\(\mu s\) pulse width. Given that EMI source has the same power and the antenna gain as in the previous two scenarios, we still have the same peak field value of 73 kV/m at 15m from the source. However, the single pulse carries all the average power of 133W per frame.

Under this scenario, we see significant failure in the timer operation (see Fig. 5.13(a)-(b)). This implies that as the repetition frequency decreases (ie. pulse period increases), impact of EMI becomes more dominant. This can also be attributed to that
Figure 5.12: (a) Incident Gaussian pulse (period of $2\mu s$ and width of $0.375\mu s$) at 15m distance from the EMI source (average power is 133W and antenna gain is 20dBi)- (b) Timer response at the output ports.

Higher number of pulses within a time frame introduces more harmonics with each having lower power.

5.2.3 Timer Performance versus Various EMI Incidence Angles

In this section, our objective is to understand the role of incident EMI angle direction on the timer operation. To do so, we illuminated the automobile with a train of Gaussian pulses (periodic with $2.0\mu s$ and pulse width of $0.375\mu s$) at carrier frequency of 1 GHz (see Fig. 5.14). Our preliminary studies suggested that an EMI source delivering 87W average power and using 20dBi antenna would generate sufficient interference to disrupt the timer.

Referring to Fig. 5.15, we show the timer response at the timer output ports, and observe that illumination from the side and rear does not affect the timer operation.
Figure 5.13: (a) Incident Gaussian pulse (period of 4$\mu$s and width of 0.75$\mu$s) at 15m distance from the EMI source (average power is 133W and antenna gain is 20dBi)- (b) Timer response at the output ports

Figure 5.14: Illustration of the timer inside an automobile illuminated from three different directions by an EMI source with 87W power and antenna gain of 20dBi
However, adverse effects are observed when the wave is incident from the front side of the car. This is very likely due to the location of the timer being in front of the engine compartment which also has perforations with the front grid.

5.3 Discussion

All the studies presented here integrated several numerical EM tools such as MoM and FEM with a circuit tool such as ADS via hybrid S-parameters. Our analysis indicated that, in addition to the EMI power level, characteristics of interfering signal such as pulse repetition rate and pulse width are as important to assess EMI/EMC performance of the system. In other words, even though electronic circuits in complex platforms such as automobiles and airplanes are highly shielded with enclosing boxes, it is still possible to upset circuits with practically available EMI power sources. Similar conclusion can also be drawn from the studies in [2]. We also note that the EMI power required to upset automobile electronics is not as high as the power used in the aforementioned scenarios since we did not consider wire bundles which represents the largest section of electronic systems in automobiles. Therefore, a fair assessment of EMI effects on RF, digital and analog circuits in complex platforms requires an extensive analysis of EM structures such as automobile body and cable bundles integrated with electronic circuits. Not only does such analysis require wisely devised hierarchial approach to break the whole problem into self-manageable components, but also theoretical tools that can combine EM structure and excitations with circuit components.
Figure 5.15: (a) Incident Gaussian pulse (period of $2\mu s$ and width of $0.375\mu s$) at 15 m distance from the EMI source for all directions (average power is 87W and antenna gain is 20dBi) - (b) Timer response at the ports (illumination from front) - (c) Timer response at the ports (illumination from side) - (d) Timer response at the ports (illumination from rear)
CHAPTER 6

CONCLUSION AND FUTURE WORK

In this dissertation, we have taken a broad multi-disciplinary approach addressing fundamental aspects of EMI on RF-Digital circuits in presence of complex platforms. Our research has primarily focused on developing theoretical tools to address mixed-signal circuits. We proposed a hybrid S-parameter matrix with the goal of extending existing port analysis to accommodate off-Board EMI. In other words, we considered external plane wave excitations as well as port (internal) sources, and proposed hybrid S-parameters for characterization of transmission line networks and mixed signal circuit systems. We introduced constant voltage sources at the ports to treat forced waves and considered induced propagating modal waves as additional entries in S-parameter matrix (hybrid S-parameters). The resulting hybrid S-matrix and voltage sources were subsequently exported to any circuit analysis tools such as HSPICE and ADS with the corresponding linear and non-linear circuit terminations at ports.

We have extensively investigated performance of digital modulation systems and an RF power amplifier subject to high power IEMI attacks. We carried out measurements and computational analysis to address both on&off-board IEMI concerns in communication systems. Our findings yielded that large power of EMI is needed to cause device level upsets in the amplifier. We also reach to the conclusion that it
practically require less power and resources to upset communication systems when
the digital modulation schemes are targeted rather than causing system level upset
by altering the device performance through high level of EMI coupling. Our anal-
ysis also suggested that nonconstant envelope digital modulation schemes are more
susceptible to EMI than its counterparts of constant envelope modulation schemes.

Hybrid S-parameter is also employed for a practical case where the timer located
inside an automobile is subject to a train of high power Gaussian pulses. In this
work, we primarily investigated how time characteristics of interfering signal affect
performance of the timer operating inside an automobile. We concluded that wide
Gaussian pulses have more impact on the performance of the timer than the narrow
pulses. We also found out that gaussian pulses with low repetition rate had been
more destructive than the pulses with higher repetition rate.

As a continuation of this work, we propose to employ hybrid S-parameters for full
system level analysis including cable bundles and mixed signals. Hybrid S-parameter
can be extracted for cable bundles following the TICE analysis. Once Hybrid S-
parameters are obtained for both cable bundles and mixed signal circuit boards, they
can be combined in a single S-matrix and exported to any circuit solver such as
ADS and HSPICE. The flow diagram in Fig. 6.1 represents the proposed approach
to full on & off – board EMI evaluation of complex systems with cable bundles and
mixed signal circuits. Since surrounding structures and PCBs are treated in EM
domain and circuit components attached to the ports are handled in circuit domain,
proposed method fully exploits the numerical techniques customized for each domain
and also allows for optimization of circuit components for better performance without
performing any further analysis for the EM structure. With the proposed method, one
might also develop circuit level protection algorithms in addition to typical shielding and protection techniques such as metallic enclosures.
APPENDIX A

OVERVIEW OF DIGITAL MODULATION SCHEMES

Digital communication systems have become very popular due to more information capacity and better quality of communication they provide. Even though there are many variants of digital modulation techniques, we explain only the four fundamental digital modulation methods, Quadrature Phase Shift Keying (QPSK), Minimum Shift Keying (MSK), Binary Phase Shift Keying (BPSK) and Quadrature Amplitude Modulation (QAM). Small variants of these techniques are employed by many applications but main features and performances primarily remain the same.

A.1 Binary Phase Shift Keying (BPSK)

BPSK represents the simplest form of phase modulation. The phase of a constant amplitude carrier is switched between 0 and $-\pi$ depending on the binaries 1 and 0, respectively. In other words, the BPSK signal takes the form

$$S(t) = A\cos(\omega_c t + \phi)$$

(A.1)

where $\phi$ takes 0 or $\pi$ depending on the binary.

Modulation and the constellation diagram for the BPSK is given in Fig 4.25.
The performance of BPSK is measured with the Bit Error Rate given by

\[ P_e = Q\left(\sqrt{\frac{2E_b}{N_0}}\right) \quad (A.2) \]

where \( E_b \) and \( N_0 \) refer to energy per bit and the white noise power in the AWGN Channel. The energy per bit is proportional to the bit duration and the signal amplitude. Considering that each symbol carries single bit, its bandwidth efficiency is 1 bit/second/Hz. Therefore, its spectral efficiency is very low compared to its counterparts. BPSK is mainly used by Deep-Space Telemetry, Wi-Fi and cable modems applications.

A.2 Quadrature Phase Shift Keying (QPSK)

QPSK is an alternative to BPSK when improved spectral efficiency is required. Stated differently, QPSK has twice the bandwidth efficiency of BPSK since it transmits 2 bits per symbol. In other words, it transmits twice as much data given the same bandwidth. Its waveforms is written as

\[ S(t) = A\cos(\omega_c t + k\frac{\pi}{4}) \quad (A.3) \]

where \( k \) is 1...4. It simply carries data by changing the phase which are \( \pi/4 \) out of phase referred as In-Phase (I) and Quadrature-Phase (Q) components. It is mainly used in Digital Video Broadcasting-Satellite (DVB-S), Code Division Multiple Access (CDMA) systems. Referring to Fig. 4.10 displaying the modulator and the constellation diagram, we notice that two orthogonal functions cosine and sine are used to upconvert each symbol after the serial to parallel converter.

To remedy shortcomings of QPSK in terms of its large transitions from one state to the another, Offset QPSK and \( \frac{\pi}{4} - QPSK \) are proposed and currently in use in
some phone systems in North America and Japan [104]. Bit Error Rate for QPSK is close to twice the rate of BPSK due to its larger bandwidth efficiency 2bit/second/Hz.

A.3 Minimum Phase Shift Keying (MSK)

QPSK described in the previous section has the disadvantage of sharp transitions leading to special efforts in RF amplifier designs to limit spectral regrowth and inefficiencies. MSK is proposed as an alternative technique to tackle this shortcoming of QPSK. Instead of using full rectangular pulses, it utilizes half sinusoids leading to sharper decay in its spectrum. MSK signal is represented as follows [104],

\[ S(t) = a_m \cos \omega_1(t) \cos \omega_c(t) - a_{m+1} \sin \omega_1(t) \sin \omega_c(t) \] (A.4)

where \( a_m \) and \( a_{m+1} \) are binary 1 and \(-1\).

Fig. 4.19 displays the modulator and the constellation diagram for the MSK. A slight variant of MSK using Gaussian pulses instead of sinusoids referred as Gaussian Minimum Shift Keying (GMSK) requiring less bandwidth but high power to achieve comparable BER performance with QPSK. It is mainly used in GSM systems. As in BPSK, it has bandwidth efficiency 1bit/second/Hz.

A.4 Quadrature Amplitude Modulation (QAM16)

As opposed to the aforementioned modulation techniques, Quadrature Amplitude Modulation (QAM) is a nonconstant amplitude modulation technique that uses both amplitude and phase to carry data. QAM16 refers to QAM with each symbol carrying four bits. This corresponds to 2bits per In-phase and 2bits per Quadrature component. Therefore, its spectral efficiency is 4bit/second/Hz which is more efficient than the aforementioned constant envelope modulation schemes. Since it transmits
more bits per symbol, its constellation points are closer to each other (see Fig. 4.31). Therefore, it is more susceptible to noise. QAM16 has many applications including Microwave Digital Radio, Wi-Fi, Digital Video Broadcasting-Cable (DVB-C) and Digital Video Broadcasting Terrestrial (DVB-T). It is mathematically represented with

\[ S(t) = a_m \cos \omega_c(t) + a_{m+1} \sin \omega_c(t) \]  

(A.5)

where \( a_m \) and \( a_{m+1} \) are pair of identifying states in the constellation of the \( i^{th} \) state.
APPENDIX B

ERROR VECTOR MAGNITUDE (EVM)

Error Vector Magnitude (EVM) is described as a measure of the difference between the measured and the ideal signal [105]. Stated differently, it measures the deviation from the ideal reference signal and represents a prevalent figure-of-merit in assessing performance of communication systems for different environments. An illustrative description of EVM is given in Fig. B.1 displaying a constellation diagram for a modulation scheme and the vector that EVM refers to.

There are also other figure-of-merits such as Bit Error Rate (BER) and eye diagrams that assess performance of communication systems. However, primary advantage of EVM over these tools is that it provides more insight into performance of the communication systems and more flexibility for designers in troubleshooting each stage of the system.

EVM is measured over time and the instantaneous EVM is the difference between the measured signal and the reference signal at any time. It is mathematically stated as

\[ EVM = |z_i - s_i| \]

where \( z_i \) and \( s_i \) refer to measured and reference symbols, respectively. Root Mean Square Error Vector Magnitude (RMS EVM) is the ratio of average error vector power.
Figure B.1: Illustration of EVM

to the ideal symbol power and computed as

$$EVM_{RMS} = \sqrt{\frac{\sum_i |z_i - s_i|^2}{\sum_i |s_i|^2}}$$  \hspace{1cm} (B.2)
BIBLIOGRAPHY


