4-21-2010

Design Considerations to Reduce Conducted and Radiated EMI

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DESIGN CONSIDERATIONS TO
REDUCE CONDUCTED AND RADIATED EMI

A Thesis

Submitted to the Faculty

of

Purdue University

by

Matthew J. Schneider

In Partial Fulfillment of the

Requirements for the Degree

of

Master of Science

May 2010

Purdue University

West Lafayette, Indiana

TABLE OF CONTENTS

Page

LIST OF TABLES………………………………………………………………… iv

LIST OF FIGURES……………………………………………………………… v

ABSTRACT……………………………………………………………………… xii

CHAPTER 1. INTRODUCTION..…………………………………………………… 1

1.1. Why is EMI Important?………………………………………………… 2
1.1.1 – Case Study 1………………………………………………………. 3
1.1.2 – Case Study 2………………………………………………………. 4
1.1.3 – Case Study 3………………………………………………………. 5
1.2. Motivation………………………………………………………………. 6
1.3. Thesis Outline……………………………………………………………7

CHAPTER 2. EMI BASICS……………………………………………………… 9

2.1. Electromagnetic Theory...………………………………………………10
2.1.1. Electrostatics…………………………………………………….. 10
2.1.2. Magnetostatics……………………………………………………14
2.1.3. Electromagnetics………………………………………………… 16
2.1.4. Propagating Electromagnetic Waves……………………………19
2.1.5. Skin Depth……………………………………………………… 22
2.2. EMC Definitions……………………………………………………… 23
2.3. EMI Measurement Requirements and Techniques………………………. 27
2.3.1. Regulatory Agency Requirements……………………………… 27
2.3.2. EMI Measurement Process………………………………………30
2.3.3. Locating Unintentional EMI…………………………………… 32
2.4. Common EMI Problems…………………………………………………34
2.4.1. Cross-Talk……………………………………………………… 35
2.4.2. Ground Loops…………………………………………………… 37
2.4.3. Antenna Effects………………………………………………….. 40
2.4.4. Spark Gaps……………………………………………………… 43

Page

2.4.5. Impedance Mismatch…………………………………………………… 44
2.4.6. Harmonics…………………………………………………………….. 46
2.5. EMI Solutions……………………………………………………………….. 47
2.5.1. Capacitive Decoupling……………………………………………... 49
2.5.2. Circuit Board Layout.…………………………………………………49
2.5.3. Ferrites..………………………………………………………………. 53
2.5.4. Impedance Matching ………………………………………………… 55
2.5.5. Cabling and Shielding………………………………………………… 57
2.5.6. Filters…………………………………………………………………. 60
2.6. EMI Simulation Software……………………………………………………… 61
2.7. Flowchart for Troubleshooting EMI……………………………………….. 63

CHAPTER 3. EMI MEASUREMENTS………………………………………………….. 66

3.1. Switching Regulator………………………………………………………… 67
3.1.1. Differential Mode Interference.……………………………………..70
3.1.2. Differential Mode Attached Loads……………………………………72
3.1.3. Common Mode Interference………………………………………….. 79
3.1.4. Common Mode Attached Loads……………………………………… 80
3.2. Digital Communication………………………………………………………... 89
3.2.1. Ribbon Cable…………………………………………………………. 90
3.2.2. Shielded Cable……………………………………………………….... 92
3.3. Coax Cables………………………………………………………………………94
3.3.1. Control…………………………………………………………………... 95
3.3.2. Damaged Braid……………………………………………………….. 97
3.3.3. Broken Braid……………………………………………………………100
3.4. Low Noise Amplifiers……………………………………………………………102
3.4.1. Control……………………………………………………………….. 103
3.4.2. Impedance Mismatch………………………………………………… 106
3.4.3. Antenna Effects…………………………………………………………108
3.4.4. Coupling ……………………………………………………………....111
3.5. Case Study: Mixed Analog and Digital Circuit Board……………………...115

CHAPTER 4. SUMMARY AND FUTURE WORK………………………………….. 127

4.1. Recommended Future Work……………………………………………….. 129

LIST OF REFERENCES……………………………………………………………………131

APPENDIX……………………………………………………………………………..133

LIST OF TABLES

Table Page

Table 2.1. Skin Depth in Copper (Sadiku, 2007)………………………………….. 23

Table 2.2. Narrowband EMI Signals (Mardiguian, 2001)…………………………. 25

Table 2.3. Broadband EMI Signals (Mardiguian, 2001)…………………………… 25

Table 2.4. Agency Requirements for Emissions (Agilent Technologies, 2000)…… 28

Table 2.5. European Requirements (Agilent Technologies, 2000)…………………. 28

Table 2.6. FCC: Class A Products at a distance of 10m (Federal Communications Commission, 2005)…………………………………………………………….. 29

Table 2.7. FCC - All devices excluding Class A at a distance of 3m (Federal Communications Commission, 2005)……………………………………………29

Table 2.8. CISPR – Quasi-Peak measurements at 10m (CISPR 22, 1997)…………29

Table 3.1. Attached Load Descriptions…………………………………………….. 70

Table 3.2. LNA Input Parameters…………………………………………………. 104

Table 3.3. Control LNA Radiated Measurements……………………………….106

Table 3.4. Unmatched LNA Radiated Measurements…………………………….. 107

Table 3.5. Antenna LNA Radiated Measurements……………………………... 111

Table 3.6. Case Study Board Partitions…………………………………………… 116

LIST OF FIGURES

Figure		Page
Figure 1.1.	Types of EMC Measurements (Agilent Technologies, 2000)……………	2
Figure 2.1.	Electric field (**E**) due to two point charges (Crowell, 2007)…………….	11
Figure 2.2.	Magnet field lines around a bar magnet (NDT Resource Center, 2008)...	14
Figure 2.3.	Propagating EM Wave (Mount Holyoke College, 2009)……………….	20
Figure 2.4.	EMI Coupling Paths (Blanchard, & Sorak, 1994)………………………	26
Figure 2.5.	Bilogrithmic Antenna (Teseq Worldwide, 2009)………………………	30
Figure 2.6.	Radiated EMI Test Setup (Agilent Technologies, 2000)………………..	31
Figure 2.7.	Capacitive Coupling Cross-talk (Intel Corporation, 2002)……………...	36
Figure 2.8.	Inductive Coupling Cross-talk (Intel Corporation, 2002)……………….	37
Figure 2.9.	Grounding Configurations (Texas Instruments, 1999)………………….	38
Figure 2.10.	Common-Mode Interference (Texas Instruments, 1999)………………..	39
Figure 2.11.	Differential-Mode Interference (Texas Instruments, 1999)……………..	39
Figure 2.12.	Monopole Antenna (Naval Air Systems Command, 1997)……………..	41
Figure 2.13.	Dipole Antenna (Naval Air Systems Command, 1997)………………….	41
Figure 2.14.	Transmission Line Model (Texas Instruments, 1994)…………………..	45
Figure 2.15.	Microstrip Transmission Line (McMahill, 2003)……………………….	46
Figure 2.16.	Cost of EMI Corrections (Freescale Semiconductor, 2005)…………….	47
Figure 2.17.	Product Development Cycle (Nageswara Rao, Venkata Ramana, Krishnamurthy, & Srinivas, 1995)…………………………………………….	48
Figure 2.18.	Decoupling Capacitor to Minimize Loop Current………………………	49
Figure 2.19.	Proper Board Zoning Layout (Texas Instruments, 1999)……………….	50
Figure 2.20.	Two-Layer PCB with “Broken” Ground (Texas Instruments, 1999)……	51
Figure 2.21.	Multi-Layer Circuit Board Layout………………………………………	52

Figure Page

Figure 2.22. Ferrite configuration for reducing CM and DM noise…………………….. 54

Figure 2.23. Impedance matching with a pull-down resistor (Texas Instruments, 1994)………………………………………………………………… 56

Figure 2.24. Impedance matching with parallel resistors (Texas Instruments, 1994)………………………………………………………………… 56

Figure 2.25. Impedance matching with a damping resistor (Texas Instruments, 1994)………………………………………………………………… 57

Figure 2.26. Ferrite used for external cabling……………………………………………59

Figure 2.27. EMI Troubleshooting Flowchart (Mardiguian, 2001)…………………….. 63

Figure 3.1. Switching Regulator Schematic…………………………………………… 68

Figure 3.2. Switching Regulator PCB………………………………………………… 69

Figure 3.3. Differential Mode No Ferrites…………………………………………….... 71

Figure 3.4. Differential Mode 100Ω Ferrites……………………………………….. 71

Figure 3.5. Differential Mode 600Ω Ferrites……………………………………….. 72

Figure 3.6. Differential Mode Load1 600Ω Ferrites……………………………………73

Figure 3.7. Differential Mode Load1 100Ω Ferrites……………………………………73

Figure 3.8. Differential Mode Load1 No Ferrites………………………………………73

Figure 3.9. Differential Mode Load2 600Ω Ferrites……………………………………74

Figure 3.10. Differential Mode Load2 100Ω Ferrites………………………………... 74

Figure 3.11. Differential Mode Load2 No Ferrites………………………………………75

Figure 3.12. Differential Mode Load3 600Ω Ferrites…………………………………....75

Figure 3.13. Differential Mode Load3 100Ω Ferrites……………………………………76

Figure 3.14. Differential Mode Load3 No Ferrites………………………………………76

Figure 3.15. Differential Mode Load4 600Ω Ferrites……………………………………77

Figure 3.16. Differential Mode Load4 100Ω Ferrites……………………………………77

Figure 3.17. Differential Mode Load4 No Ferrites……………………………………....77

Figure 3.18. Differential Mode Load5 600Ω Ferrites………………………………... 78

Figure 3.19. Differential Mode Load5 100Ω Ferrites………………………………... 78

Figure 3.20. Differential Mode Load5 No Ferrites………………………………………79

Figure 3.21. Common Mode No Ferrites……………………………………………….. 80

Figure Page
Figure 3.22. Common Mode Load1 600Ω Ferrites………………………………….. 81
Figure 3.23. Common Mode Load1 100Ω Ferrites………………………………….. 81
Figure 3.24. Common Mode Load1 No Ferrites…………………………………….. 82
Figure 3.25. Common Mode Load2 600Ω Ferrites………………………………….. 82
Figure 3.26. Common Mode Load2 100Ω Ferrites………………………………….. 83
Figure 3.27. Common Mode Load2 No Ferrites…………………………………….. 83
Figure 3.28. Common Mode Load3 600Ω Ferrites………………………………….. 84
Figure 3.29. Common Mode Load3 100Ω Ferrites………………………………….. 84
Figure 3.30. Common Mode Load3 No Ferrites…………………………………….. 84
Figure 3.31. Common Mode Load4 600Ω Ferrites………………………………….. 85
Figure 3.32. Common Mode Load4 100Ω Ferrites………………………………….. 85
Figure 3.33. Common Mode Load4 No Ferrites…………………………………….. 86
Figure 3.34. Common Mode Load5 600Ω Ferrites………………………………….. 86
Figure 3.35. Common Mode Load5 100Ω Ferrites………………………………….. 87
Figure 3.36. Common Mode Load5 No Ferrites…………………………………….. 87
Figure 3.37. Differential Mode Reduction in Radiated Emissions……………………88
Figure 3.38. Common Mode Reduction in Radiated Emissions………………………88
Figure 3.39. Ribbon Cable with No Ferrites…………………………………………. 91
Figure 3.40. Ribbon Cable with Toroidal Ferrites…………………………………… 91
Figure 3.41. Shielded Cable with No Ferrites……………………………………….. 92
Figure 3.42. Shielded Cable with Toroidal Ferrites…………………………………... 93
Figure 3.43. Coaxial Cable (CommScope Properties, 2007)………………………… 94
Figure 3.44. Good Cable – Radiated at Load (Input Signal Frequency = 0.25GHz)…………………………………………………………. 96
Figure 3.45. Good Cable – Radiated at Load (Input Signal Frequency = 1.0GHz)……………………………………………………………96
Figure 3.46. Good Cable – Radiated at Load (Input Signal Frequency = 1.75GHz)…………………………………………………………. 96
Figure 3.47. Damaged Cable – Radiated at Load (Input Signal Frequency = 0.25GHz)…………………………………………………………. 98

Figure Page

Figure 3.48. Damaged Cable – Radiated at Load (Input Signal Frequency = 1.0GHz)…… 98

Figure 3.49. Damaged Cable – Radiated at Load (Input Signal Frequency = 1.75GHz)…… 98

Figure 3.50. Damaged Cable – Radiated at Generator (Input Signal Frequency = 0.25GHz)…… 99

Figure 3.51. Damaged Cable – Radiated at Generator (Input Signal Frequency = 1.0GHz)…… 99

Figure 3.52. Damaged Cable – Radiated at Generator (Input Signal Frequency = 1.75GHz)…… 100

Figure 3.53. Broken Cable – Radiated at Load (Input Signal Frequency = 0.25GHz)…… 101

Figure 3.54. Broken Cable – Radiated at Load (Input Signal Frequency = 1.0GHz)…… 101

Figure 3.55. Broken Cable – Radiated at Load (Input Signal Frequency = 1.75GHz)…… 102

Figure 3.56. Control LNA Schematic…… 103

Figure 3.57. Control LNA Layout…… 104

Figure 3.58. Control Normal Operating Conditions…… 105

Figure 3.59. Control in Compression…… 105

Figure 3.60. Unmatched Normal Operating Conditions…… 107

Figure 3.61. Unmatched LNA in Compression…… 107

Figure 3.62. Antenna LNA Schematic…… 108

Figure 3.63. Antenna LNA Layout…… 109

Figure 3.64. Antenna Effects Normal Operating Conditions…… 110

Figure 3.65. Antenna Effects in Compression…… 110

Figure 3.66. Coupling LNA Schematic…… 112

Figure 3.67. Coupling LNA Layout…… 112

Figure 3.68. Conducted EMI Normal Operation…… 113

Figure 3.69. Conducted EMI in Compression…… 113

Figure 3.70. LNA Radiated Emissions Comparison…… 114

Figure Page

Figure 3.71. Section 1 No Ferrites.. 117
Figure 3.72. Section 1 No Ferrites Standby... 117
Figure 3.73. Section 2 No Ferrites.. 117
Figure 3.74. Section 2 No Ferrites Standby... 117
Figure 3.75. Section 3 No Ferrites.. 117
Figure 3.76. Section 3 No Ferrites Standby... 117
Figure 3.77. Section 4 No Ferrites.. 118
Figure 3.78. Section 4 No Ferrites Standby... 118
Figure 3.79. Section 5 No Ferrites.. 118
Figure 3.80. Section 5 No Ferrites Standby... 118
Figure 3.81. Section 6 No Ferrites.. 118
Figure 3.82. Section 6 No Ferrites Standby... 118
Figure 3.83. Section 1 Digital Only... 119
Figure 3.84. Section 1 Digital Only Standby.. 119
Figure 3.85. Section 2 Digital Only... 119
Figure 3.86. Section 2 Digital Only Standby.. 119
Figure 3.87. Section 3 Digital Only... 120
Figure 3.88. Section 3 Digital Only Standby.. 120
Figure 3.89. Section 4 Digital Only... 120
Figure 3.90. Section 4 Digital Only Standby.. 120
Figure 3.91. Section 5 Digital Only... 120
Figure 3.92. Section 5 Digital Only Standby.. 120
Figure 3.93. Section 6 Digital Only... 121
Figure 3.94. Section 6 Digital Only Standby.. 121
Figure 3.95. Section 1 A&D Ferrites... 121
Figure 3.96. Section 1 A&D Ferrites Standby.. 121
Figure 3.97. Section 2 A&D Ferrites... 122
Figure 3.98. Section 2 A&D Ferrites Standby.. 122
Figure 3.99. Section 3 A&D Ferrites... 122

Figure Page

Figure 3.100. Section 3 A&D Ferrites Standby………………………………………. 122

Figure 3.101. Section 4 A&D Ferrites…………………………………………….... 122

Figure 3.102. Section 4 A&D Ferrites Standby………………………………………. 122

Figure 3.103. Section 5 A&D Ferrites…………………………………………….... 123

Figure 3.104. Section 5 A&D Ferrites Standby………………………………………. 123

Figure 3.105. Section 6 A&D Ferrites…………………………………………….... 123

Figure 3.106. Section 6 A&D Ferrites Standby………………………………………. 123

Figure 3.107. Section 1 600Ω Ferrites…………………………………………….... 124

Figure 3.108. Section 1 600Ω Ferrites Standby………………………………………. 124

Figure 3.109. Section 2 600Ω Ferrites…………………………………………….... 124

Figure 3.110. Section 2 600Ω Ferrites Standby………………………………………. 124

Figure 3.111. Section 3 600Ω Ferrites…………………………………………….... 124

Figure 3.112. Section 3 600Ω Ferrites Standby………………………………………. 124

Figure 3.113. Section 4 600Ω Ferrites…………………………………………….... 125

Figure 3.114. Section 4 600Ω Ferrites Standby………………………………………. 125

Figure 3.115. Section 5 600Ω Ferrites…………………………………………….... 125

Figure 3.116. Section 5 600Ω Ferrites Standby………………………………………. 125

Figure 3.117. Section 6 600Ω Ferrites…………………………………………….... 125

Figure 3.118. Section 6 600Ω Ferrites Standby………………………………………. 125

Appendix Figure

Figure A.1. Common Mode 100Ω Ferrites…………………………………………. 133

Figure A.2. Common Mode 600Ω Ferrites…………………………………………. 133

Figure A.3. Good Cable – Radiated at Generator (Input Signal Frequency = 0.25GHz)…………………………………………………… 134

Figure A.4. Good Cable – Radiated at Generator (Input Signal Frequency = 1.0GHz)……………………………………………………... 134

Figure A.5. Good Cable – Radiated at Generator (Input Signal Frequency = 1.75GHz)…………………………………………………… 134

Figure A.6. Good Cable – Radiated at Middle (Input Signal Frequency = 0.25GHz)…………………………………………………… 135

Appendix Figure Page

Figure A.7. Good Cable – Radiated at Middle (Input Signal Frequency = 1.0GHz)………………………………………………………….. 135

Figure A.8. Good Cable – Radiated at Middle (Input Signal Frequency = 1.75GHz)………………………………………………………… 135

Figure A.9. Damaged Cable – Radiated at Middle (Input Signal Frequency = 0.25GHz)………………………………………………………… 136

Figure A.10. Damaged Cable – Radiated at Middle (Input Signal Frequency = 1.0GHz)………………………………………………………….. 136

Figure A.11. Damaged Cable – Radiated at Middle (Input Signal Frequency = 1.75GHz)………………………………………………………… 136

Figure A.12. Broken Cable – Radiated at Middle (Input Signal Frequency = 0.25GHz)………………………………………………………… 137

Figure A.13. Broken Cable – Radiated at Middle (Input Signal Frequency = 1.0GHz)………………………………………………………….. 137

Figure A.14. Broken Cable – Radiated at Middle (Input Signal Frequency = 1.75GHz)………………………………………………………… 137

Figure A.15. Broken Cable – Radiated at Generator (Input Signal Frequency = 0.25GHz)………………………………………………………… 138

Figure A.16. Broken Cable – Radiated at Generator (Input Signal Frequency = 1.0GHz)………………………………………………………….. 138

Figure A.17. Broken Cable – Radiated at Generator (Input Signal Frequency = 1.75GHz)………………………………………………………… 138

ABSTRACT

Schneider, Matthew J. M.S., Purdue University, May 2010. Design Considerations to Reduce Conducted and Radiated EMI. Major Professor: John P. Denton.

This work evaluates the effects of conducted and radiated EMI due to improperly designed circuits or printed circuit board (PCB) layouts. Common circuits used in a wide range of electronic devices are evaluated using both good and poor design techniques. Additional evaluations are performed to measure the effects of EMI on cabling used for digital communication or RF purposes. Finally, a consumer product is evaluated which must pass regulatory agency requirements in order to be sold in the consumer market. Potential solutions to common EMI problems are then implemented and discussed. This work will look at individual circuits and evaluate design flaws that contribute the largest amount of radiated emissions. Avoiding poor circuit designs and PCB layout mistakes reduces the amount of radiated emissions from any product. This work discusses the theory of electromagnetics and how it pertains to EMI events, then addresses the practical problems associated with EMI using common circuits to illustrate EMI effects. It is important to understand why EMI related effects are becoming more critical, and to show why EMI problems should be corrected within the design phase.

CHAPTER 1. INTRODUCTION

Electromagnetic interference (EMI) is everywhere and unavoidable. It exists in nature and is also manmade. EMI is the presence of unwanted electromagnetic energy which has the potential to cause disturbances in electronic devices. Sources in nature can come from lightning and electrostatic discharges; while man-made EMI can originate from motors, power lines, fluorescent bulbs, and many other places (Levitt, 2001). Any object which has a time-varying electric or magnetic field can be a potential source of EMI. All electronic devices have the potential to be sources of EMI by generating conducted or radiating emissions, along with being victims by accepting EMI from other sources. The amount of EMI a device contributes is referred to as conducted or radiated EMI, and the amount of EMI the device is able to withstand is referred to as EMI susceptibility. In an ideal world the perfect device would not conduct or radiate any EMI and would be perfectly immune to EMI susceptibility, however due to the way electronics work this is an impossible task to achieve. All devices will conduct or radiate some EMI; the amount must be regulated to below a threshold set by the FCC in Section15 of its regulations.

Any electronic device which is to be bought or sold must be certified that it meets electromagnetic compatibility (EMC) requirements. To meet EMC requirements the device must be tested for conducted and radiated emissions along with conducted and radiated susceptibility. Figure 1.1 shows an example of the two types of EMC measurements.

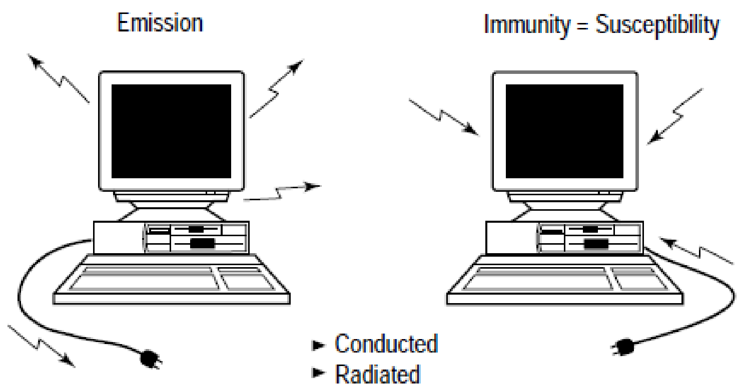


Figure 1.1. Types of EMC Measurements (Agilent Technologies, 2000).

EMC requirements vary depending on which country the product will be bought or sold in, and how the product is classified whether that is a consumer, medical, or military product. Many times during the design phase EMC is completely forgotten or disregarded until the final stage of development. This lack of consideration can cause added expense and delay product time to market. EMC must be considered in all aspects of the design from the initial drawings to layout and packaging.

1.1. Why is EMI Important?

With consumer electronics continuously on the rise, the effects of EMI on these devices must be taken into consideration. It would be very annoying if a person walking down the sidewalk talking on their cell phone caused interference to someone else listening to an audio device, or vice versa. The effects of EMI can be heard from cell phones coupling with computer speakers, and can be seen by static noise on analog televisions. If the magnitude of the EM field is large enough it has the potential to destroy sensitive electronic equipment. The susceptibility of electronic equipment must be hardened to increase immunity, along with proper design techniques used to prevent conducted and radiated emissions.

There have been multiple cases, some more serious than others, of electronic equipment malfunctioning due to EMI. There are very strict regulations for EMI in medical equipment to insure proper device operation during critical situations. Other disturbances from EMI, such as noise in an analog television, can be minor annoyances. A few case studies are presented in order to demonstrate the effects of EMI.

1.1.1. Case Study 1

Assisted listening devices (ALDs), such as hearing aids, are designed to amplify audio signals so that a person with low hearing can communicate with another person more effectively. Hearing aids have two modes of operation. The first is microphone mode, and the second mode uses a small telecoil. The telecoil is a small coil of wire which is sensitive to changes in magnetic fields. This was originally developed to help individuals with low hearing use the telephone, and is currently used for assisted learning systems which use special headsets or receivers. With the rise of digital consumer electronics more devices are causing interference with the telecoil. Magnetic fields in the hearing range of 20Hz-20kHz directly couple with the telecoil and are amplified in the initial stages of the hearing aid. Other higher frequency signals can also couple with the telecoil. Although these frequencies are higher than a person can hear, non-linear components such as mixers in the demodulation portion of the receiver can generate lower frequencies which result in noise. Two methods are used to reduce EMI susceptibility in hearing aids. The first method is to use shielding techniques to prevent EMI from reaching sensitive components. The second method uses a balancing technique by distributing EMI equally on both sides of a sensitive circuit. Since there will be no difference in potential, the noise produced by EMI is ignored and only the good signal is demodulated. The second scheme has been generally accepted better than the first. EMI in hearing aids is a growing concern and the problem is still in the process of being addressed (Levitt, 2001).

1.1.2. Case Study 2

An aeronautical radio navigation station was built to navigate airplanes within an 80km radius around the station. The portion of the station responsible for communications consisted of a signal transmitting module, a T-shaped antenna, and radial ground grids. The output power of the transmitting module was 500W, and roughly 76% of the power radiated through the antenna. Located at 30m and 35m away were 11.4kV power lines and telephone wires. The power lines were run to a transformer which dropped the voltage to 115/230V and fed power to the station. When the station was operational there were many problems that prevented the station from operating normally. An auto voltage regulator could not automatically adjust the output voltage to the transmitter of the pilot signal, the air-conditioning unit would not operate normally, a telephone could not dial out but was able to receive calls, and arcing traces were found on the circuit board of the emergency lighting unit. Measurements were made of the stations main power input when the station was broadcasting and the carrier signal was found to be coupling onto the power grid and riding on top of the 115V input to the station.

Electromagnetic energy from the antenna was coupling onto the outside power lines and getting fed back into the station which was causing multiple problems. In order to reduce EMI to a reasonable level the overhead power lines were routed with underground cables, a multipoint grounding system was used, and the telephone set was shielded with a copper box. After the fixes the radio station operated normally (Chia-Hao, & Chang-Fa, 2006).

1.1.3. Case Study 3

A relatively new discovery in EMI concerns is the impact it has on pacemakers and defibrillators. A study was done using headphones and placing them in close proximity to different heart devices. Earphones can have strong magnets in them which will be surrounded by a magnetic field, the impact this field had on heart devices was to be studied. Sixty patients were chosen for a study which placed 8 different types of ear-bud or clip-on headphones close to a heart device implanted in their chest. It was found that EMI had an impact on 14 of the individual's heart devices, or 23% of the participants in the study. The problem is more serious for individuals who wear defibrillators because the EMI completely disabled some of the devices. It was found that at a distance of 3cm above the person's skin that there were no problems. Someone with a heart device should avoid laying headphones on their chest, or letting someone lean against their chest that is using headphones. In the United States approximately 250,000 people are fitted with a pacemaker, and 125,000 receive defibrillators used to shock the heart back into beating a constant rate. Any interference with the operation of these devices could have a fatal impact on the individual. This study is still being evaluated and more evidence is needed to determine the true impact on heart devices (Patrick, 2008).

As can be seen with the previous case studies the effects of EMI can impact a wide range of areas. Depending on the application, the effects of EMI can range from mere inconveniences, to life threatening situations. It is clearly seen that the effects of EMI are generally unwanted and negative.

1.2. Motivation

Motivation for this work stems from current work experiences. Post graduation and being introduced into a new work environment, EMC was the least of concerns. Completing designs on time and meeting product deadlines were of top priority. It wasn't until it came to getting products to market, and going through FCC testing when, it was realized how important designing for EMC became. After going through testing multiple times and gaining firsthand experience and knowledge about EMC requirements, it was realized that multiple design changes could have been done earlier to prevent some of the issues. As most designs go, EMC was forgotten about until it came time to pass FCC requirements. When the product failed testing it came down to last minute changes in order to get the product in accordance with regulations. Through this experience it became apparent how underprepared students are for this scenario.

Currently in coursework students learn about EMI and some of the effects, but most material is brief, and there are no labs to support the lecture material. Articles have been previously published suggesting that EMI related studies should be implemented in the Engineering and Technology curriculum. D. Blanchard, from Purdue University

Calumet, published a paper in 1994 titled “EMI is Everybody’s Concern. Where is it in the Engineering and Technology Curriculum?” that outlined an entire course which could be used to teach students about EMI. In this paper, D. Blanchard suggests that the course cover a range of issues such as EMI fundamentals, source detection, receptors and radiators, regulations and standards and measurements techniques (Blanchard, & Sorak, 1994). A second paper proposed in 1995 by G.K. Deb titled “Relevance of EMC Education in Undergraduate Course” suggested another course intended for students interested in EMC Engineering (Deb, 1995). The outline of this course would cover EMI problems and hardening techniques, analysis techniques, computational electromagnetics, methods of EMI measurements, various standards and specifications used in the international levels, and instrumentation. The goal of both papers was to get EMC information to students before graduation. These papers were published roughly 15 years ago, and since then no dedicated EMC course has been implemented for undergraduate students. EMI problems are not going to disappear, and EMC is becoming more critical as consumers use more electronic devices in their everyday lives.

1.3. Thesis Outline

This thesis will cover a broad range of EMI topics beginning with electromagnetic theory. Basic fundamentals of point charges, Maxwell’s equations, and finally transverse electromagnetic waves will all be examined. EMI definitions will be covered and measurement techniques analyzed. Sources of EMI will be discussed such as conducted EMI, transients, ground loops, common and differential mode currents, coupling, and cross-talk. Major contributors to radiated EMI will be identified, and design solutions proposed to reduce the magnitude of the radiated fields. These proposed solutions will primarily focus on PCB layout design, but will also include shielding, coupling paths, and ground loops. A flowchart is also presented in order to identify EMI problems and reduce emissions.

For the major problems which contribute to radiated EMI, specific circuits are designed which intentionally exacerbate the amount of emissions. This is done to classify which circuits contribute the largest amount of EMI emissions from any product which is to be sold on the market. Using the circuits designed to exacerbate radiated EMI, comparisons will be made to a good design to identify the increase in radiated EMI. By identifying which circuit issues contribute the most to the amount of radiated EMI, design considerations can be implemented which will yield the highest reduction in emissions for almost any product. Finally, a case study is performed on a prototype circuit board in order to prepare for success during EMI testing. This prototype circuit board incorporates many of the corrected circuit designs which were previously designed to intentionally increase emissions. The overall goal of the thesis is to educate circuit and system designers with an awareness of design related issues which contribute to EMI, and to show why EMI problems should be corrected early in the design phase rather than patched at the end.

CHAPTER 2. EMI BASICS

In order to provide a good understanding of EMI, this chapter will begin by introducing electromagnetic theory. Starting with electrostatics and magnetostatics, and then introducing time varying electromagnetic fields. The coupling between the electric and magnetic fields will be explained using Maxwell's equations, and an introduction to propagating electromagnetic waves relating to EMI will be provided. Finally, impedance matching techniques will be discussed, and how it relates to EMI.

Definitions specifically relating to EMI will be explained prior to getting too deep into EMI issues. These definitions will explain EMI concepts in terms of separate domains, span of frequency spectrum, and measurement quantities. Measurement methods and specific techniques will be introduced in accordance with the Federal Communications Commission (FCC) Section 15, and the European International Special Committee on Radio Interference (CISPR) regulations. These methods will include environmental conditions, along with EMI test equipment and antennas.

An overview of the fundamental problems which contribute to EMI will be introduced. These problems will be separated into two categories: conducted and radiated EMI. Problems which will be discussed are common-mode and differential mode noise, capacitive and inductive coupling, cross-talk, and spark gaps. This chapter will then present a section on solutions to the discussed problems.

Finally a flowchart for troubleshooting EMI problems during the testing phase will be presented. This flowchart will provide a step-by-step method to determine EMI problems, and reduce the amount of guessing and frustration which commonly plagues EMC test engineers.

2.1. Electromagnetic Theory

This section presents a brief introduction to electromagnetic theory. Electromagnetic interference is governed by the same laws and principals that are used today for all electronic devices, however it is unwanted. This section will present the very basics of EMI, and why it behaves in the manner it does.

2.1.1. Electrostatics

The source of an electric field is an electric charge. Electric charges can have either a positive or negative charge. Point charges exist at the atomic level and are associated with electrons and protons. An electron is negatively charged where the proton is positively charged. Both the electron and proton have the same magnitude of charge equal to *e* (1.6 x 10^{-19} Coulombs).

$$q_e = -e \qquad \text{Charge of single electron} \tag{2.1}$$

$$q_p = e \qquad \text{Charge of single proton} \tag{2.2}$$

A single point charge will have an associated electric field. A positively charged particle is a *source* of an electric field, and a negatively charged particle is a *sink* of an electric field. The source of an electric field means that the field extends outward from the particle, whereas a sink terminates an electric field. Figure 2.1 shows an example of the electric field due to two point charges.

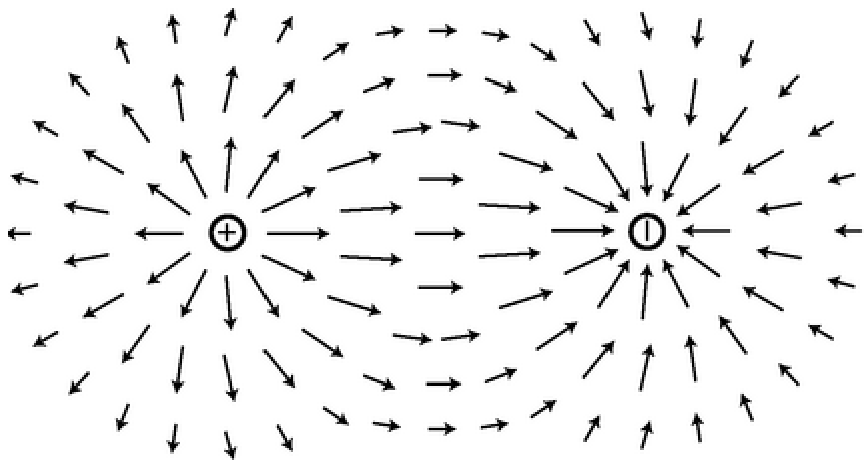


Figure 2.1. Electric field (**E**) due to two point charges (Crowell, 2007).

The intensity of an electric field associated with a point charge is given by:

$$\mathbf{E} = \frac{q}{4\pi\varepsilon R^2}\hat{\mathbf{r}} \quad (\mathrm{V/m}) \tag{2.3}$$

From equation 2.3, $\hat{\mathbf{r}}$ is a unit vector in the direction of the electric field, q is the magnitude of the point charge, R is the distance from the observation point to the charge, and ε is the permittivity of the material. The permittivity of a material describes the interaction between the electric field and the material surrounding it. More specifically the permittivity describes how well an electric field is able to travel through the material, the permittivity for free space is given by equation 2.4.

$$\varepsilon = \varepsilon_0 = 8.854 \times 10^{-12} \text{ (F/m)} \tag{2.4}$$

The permittivity of free space has a relative permittivity ($\varepsilon_r = 1$). Materials such as glass, water, and different substrates will all exhibit different relative permittivities, and all impact the electric field differently. For these materials the permittivity can be determined using equation 2.5.

$$\varepsilon = \varepsilon_0 \varepsilon_r \tag{2.5}$$

In addition to point charges, line charges, surface charges, and volume charges exist as different charge distributions. Charge density for line charges, surface charges, and volume charges can be denoted by ρ_L, ρ_S and ρ_V, respectively. The electric field due to each charge distribution can be solved for by summing a small portion of the line (*dl*), surface (*ds*), or volume (*dv*), over the entire line, surface, or volume. Solving for the electric field based on these charge distributions it can be found that:

$$\mathbf{E} = \int_{\boldsymbol{L}} \frac{\rho_L dl}{4\pi\varepsilon R^2}\,\hat{\mathbf{r}} \qquad \text{Line Charge} \tag{2.6}$$

$$\mathbf{E} = \int_{\boldsymbol{s}} \frac{\rho_s ds}{4\pi\varepsilon R^2}\,\hat{\mathbf{r}} \qquad \text{Surface Charge} \tag{2.7}$$

$$\mathbf{E} = \int_{\boldsymbol{v}} \frac{\rho_v dv}{4\pi\varepsilon R^2}\,\hat{\mathbf{r}} \qquad \text{Volume Charge} \tag{2.8}$$

An important concept with charge distributions is the superposition principal. The superposition principal allows for a collection of point, line, surface, and volume charges to be summed together in order to solve for the net electric field. The sum of the net charges into the electric field is what could eventually become an EMI problem. By expanding the unit vector from equation 2.3, and summing a collection of point charges from 1 to N the expression becomes:

$$\mathbf{E} = \frac{1}{4\pi\varepsilon}\sum_{k=1}^{N} \frac{q_k(\mathbf{r}-\mathbf{r}_k)}{|\mathbf{r}-\mathbf{r}_k|^3} \tag{2.9}$$

The electric potential can also be solved from a charge distribution. The electric potential is a scalar unit, and is expressed in volts. For a specific volume the electric potential is given by:

$$V = \int_v \frac{\rho_v dv}{4\pi\varepsilon R} \quad \text{(V)} \tag{2.10}$$

Along with the electric field intensity, there is an associated electric flux density. Like the electric field intensity, the electric flux is also a vector. The electric flux density is commonly referred to as the electric displacement field, the two quantities are identical and can be interchanged. The flux density is proportional to the electric field intensity, and is related via the permittivity of the material. The electric flux density is expressed mathmatically as:

$$\mathbf{D} = \varepsilon\mathbf{E} \quad \text{(C/m}^2\text{)} \tag{2.11}$$

The total electric flux through a surface can now be calculated as:

$$\Psi = \int_S \mathbf{D} \cdot \mathrm{d}\mathbf{S} \quad \text{(C)} \tag{2.12}$$

There are two fields which must be looked at when discussing EMI. The electric field, which was just discussed, and a magnetic field. From the equations presented so far in this chapter, it can be seen how different charge distributions can produce an electric field, and a volume charge can result in a voltage potential. The electric field is the dual to the magnetic field. The magnetic field is produced by current distributions and has some similarities to the electric field. Both the electric and magnetic fields couple to create EMI issues, and both must be understood. Where the electric field is capable of inducing a voltage in a wire, a magnetic field is capable of inducing a current.

2.1.2. Magnetostatics

Every magnet has a north and south pole. A magnetic field will extend from the north pole, loop back on itself, and re-enter the south pole. Unlike the electric field which terminates on a negative charge, the magnetic field is more like a rubber band, and terminates on itself. Figure 2.2 shows a bar magnet with an associated magnetic field (**H**).

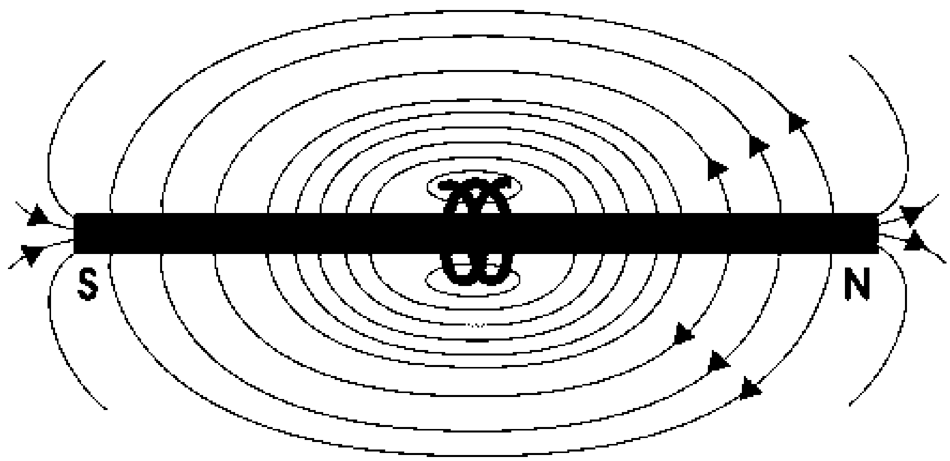


Figure 2.2. Magnet field lines around a bar magnet (NDT Resource Center, 2008).

Like the electric field which had a specific intensity, and density, the magnetic field has similar quantities. The magnetic field intensity (H) is is produced by a current, and commonly denoted in vector notation. The differential magnetic field intensity is denoted as:

$$d\mathbf{H} = \frac{Id\boldsymbol{l}\times\hat{\mathbf{r}}}{4\pi R^2} = \frac{Id\boldsymbol{l}\times\mathbf{R}}{4\pi R^3} \quad (\mathrm{A/m}) \qquad (2.13)$$

Where I is the current, $\hat{\mathbf{r}}$ is a unit vector, R = |**R**|, and $\hat{\mathbf{r}} = \mathbf{R}/R$. The magnetic field intensity is measured in units of amps per meter. The density of the magnetic field (**B**) is related to the magnetic field intensity through the permeability of the material the fields are in.

$$\mathbf{B} = \mu\mathbf{H} \tag{2.14}$$

Not to be confused with the permittivity of a material, which impacts the electric field, the permeability of a material influences the magnetic field. The permeability of a material describes how well a magnetic field is able to travel through the material. For free space the permeability is equal to:

$$\mu_0 = 4\pi \text{ x } 10^{-7} \text{ (H/m)} \tag{2.15}$$

Most materials are non-magnetic and have a relative permeability ($\mu_r = 1$), however some materials such as iron, steel, and nickel have high permeability's and allow the magnetic field to travel through them much easier than free space. For these materials the permeability is determined using equation 2.16.

$$\mu = \mu_0\mu_r \tag{2.16}$$

Similar to electric charges with different distributions, currents can also have multiple distributions. These distributions are commonly in the form of a line current (**I**), surface current (**K**), or volume current (**J**). The magnetic field intensity can be solved for each distribution from:

$$\mathbf{H} = \int_{\boldsymbol{L}} \frac{\mathrm{I}d\boldsymbol{l} \times \hat{\mathbf{r}}}{4\pi \mathrm{R}^2} \qquad \text{Line Current} \tag{2.17}$$

$$\mathbf{H} = \int_{\boldsymbol{s}} \frac{\mathbf{K}dS \times \hat{\mathbf{r}}}{4\pi \mathrm{R}^2} \qquad \text{Surface Current} \tag{2.18}$$

$$\mathbf{H} = \int_{\boldsymbol{v}} \frac{\mathbf{J}dv \times \hat{\mathbf{r}}}{4\pi \mathrm{R}^2} \qquad \text{Volume Current} \tag{2.19}$$

A simple way to visualize the cross product of two vectors, and determine the direction of a magnetic field, is to use the right hand rule. By pointing the thumb on your right hand in the direction of conventional current, and wrapping your fingers around the wire, the direction of the magnetic field is in the same direction as your fingers.

Again, similar to the electric potential, the magnetic field has an associated potential. The magnetic potential is a vector. For a specific volume the magnetic potential can be defined as:

$$\mathbf{A} = \int_v \frac{\mu \mathbf{J} dv}{4\pi R} \qquad (2.20)$$

2.1.3. Electromagnetics

When discussing electrostatics and magnetostatics, the electric field and magnetic field are independent of one another. Once the fields are no longer static and vary with time, the fields couple to each other. A time-varying electric field will generate a magnetic field, and a time-varying magnetic field will generate an electric field. Using laws determined by Gauss, Faraday, and Ampere, Maxwell came up with a set of four differential equations which describe electromagnetic fields.

Gauss's law states that "*The electric flux through any closed surface is proportional to the enclosed electric charge.*" Maxwell expressed this law mathmatically as:

$$\oint_s \mathbf{E} \cdot d\mathbf{A} = \frac{\mathrm{Q_{enclosed}}}{\varepsilon} \qquad \text{Integral Form} \qquad (2.21)$$

$$\nabla \cdot \mathbf{E} = \frac{\rho}{\varepsilon} \qquad \text{Differential Form} \qquad (2.22)$$

Gauss also has a law for magnetism which states that magnetic charges do not exist. This means magnetic monopole cannot exist. The magnetic field must be continious, and only magnetic dipoles can exist. This law is expressed as:

$$\oint_s \mathbf{B} \cdot d\mathbf{A} = 0 \qquad \text{Integral Form} \qquad (2.22)$$

$$\nabla \cdot \mathbf{B} = 0 \qquad \text{Differential Form} \qquad (2.23)$$

Faraday's law for induction is commonly used in generators and transformers. Faradays law states *"The induced EMF in any closed circuit is equal to the time rate of change of the magnetic flux through the circuit."* Lenz also contributed to Faraday's law by stating *"The current in the circuit is always in such a direction as to oppose the change of magnetic flux that produced it."* Lenz's law is the reason for the negative sign in equations 2.4 and 2.5. For a stationary surface S, Faraday's law as expressed by Maxwell is:

$$\oint_C \mathbf{E} \cdot d\boldsymbol{l} = -\int_s \frac{\partial \mathbf{B}}{\partial t} \cdot d\boldsymbol{s} \qquad \text{Integral Form} \qquad (2.24)$$

$$\nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t} \qquad \text{Differential Form} \qquad (2.25)$$

Amperes law is used to relate the magnetic field around a closed loop, to the magnitude of the current passing through the loop. Amperes original equation was modified by Maxwell to compensate for the displacement current, or electric displacement field, which also has an associated magnetic field. Maxwell's modified equations describing Faradays law are as follows:

$$\oint_C \mathbf{H} \cdot d\boldsymbol{l} = \int_s \left(\mathbf{J} + \frac{\partial \mathbf{D}}{\partial t}\right) \cdot d\boldsymbol{s} \qquad \text{Integral Form} \qquad (2.26)$$

$$\nabla \times \mathbf{H} = \mathbf{J} + \frac{\partial \mathbf{D}}{\partial t} \qquad \text{Differential Form} \qquad (2.27)$$

Using Maxwells equations any structure can be analyzed to solve for the corresponding E and H fields. Commonly when discussing EMI, simple shapes such as a straight wires or loops have an antenna effect which radiates or couples interference to the system. Using the previous equations from Maxwell, the E and H fields for a straight wire and loop can be solved. Since the derivations can be quite tedious only the solutions will be presented. For both examples, the expression for the fields vary depending if you are in the near or far field. The requirements to be considered in the near or far field are:

$$\mathrm{D} < \frac{\lambda}{2\pi} \qquad \text{Near Field Requirement} \qquad (2.28)$$

$$\mathrm{D} > \frac{\lambda}{2\pi} \qquad \text{Field Requirement} \qquad (2.29)$$

Where D is the distance to the observation point, and lambda is the wavelength of the signal. The wavelength can be calculated from frequency using the following:

$$\lambda = \frac{c}{\mathrm{f}} \qquad (2.30)$$

Where *c* is the speed of light, and f is the frequency of the signal.

For a straight wire the field E and H intensities are found to be (Mardiguian, 2001):

$$\mathrm{E} = \frac{\mathrm{Z_0} \mathrm{I} l \lambda}{8\pi^2 \mathrm{D}^3} \qquad \text{Near E-Field} \qquad (2.31)$$

$$\mathrm{H} = \frac{\mathrm{I} l}{4\pi \mathrm{D}^2} \qquad \text{Near H-Field} \qquad (2.32)$$

$$\mathrm{E} = \frac{\mathrm{Z_0} \mathrm{I} l}{2\lambda \mathrm{D}} \qquad \text{Far E-Field} \qquad (2.33)$$

$$\mathrm{H} = \frac{\mathrm{I} l}{2\lambda \mathrm{D}} \qquad \text{Far H-Field} \qquad (2.34)$$

Where Z_o is the free space impedance (377Ω), I is the current, *l* is the wire length, λ is the wavelength, and D is the distance to the observation point.

For a loop the near and far field requirements are the same. The E and H field intensities for the loop can be determined from (Mardiguian, 2001):

$$E = \frac{Z_o IA}{2\lambda D^2} \quad \text{Near E-Field} \quad (2.35)$$

$$H = \frac{IA}{4\pi D^3} \quad \text{Near H-Field} \quad (2.36)$$

$$E = \frac{Z_o \pi IA}{\lambda^2 D} \quad \text{Far E-Field} \quad (2.37)$$

$$H = \frac{\pi IA}{\lambda^2 D} \quad \text{Far H-Field} \quad (2.38)$$

It can be seen from the equations for the far field both E and H fall off as 1/D. In the near field E and H will fall fall off as $1/D^2$ or $1/D^3$ depending on if the source is a straight wire or loop. These equations show how the fields behave differently, and should help predict the amount of radiated emissions due to the antenna effect of wires or printed traces.

2.1.4. Propagating Electromagnetic Waves

Now that the electric and magnetic field intensities and coupling have been explained, propagating electromagnetic waves can be discussed. Electromagnetic waves are what constitute electromagnetic radiation, and are the fundamental source of EMI problems. Electromagnetic radiation is based on a particle model. The specific energy of an EM wave is quantized, meaning it consists of discrete packets of energy. These packets of energy are transported by photons. Photons are absorbed and emitted due to changing energy levels of electrons, and will have an energy level defined by the Plank-Einstein equation shown in equation 2.39 (Sadiku, 2007).

Energy = hf Plank – Einstein Equation (2.39)

Where h = 6.626 x 10^{-23} J·s and is Plank's constant, and f is the frequency of the photon. Maxwell predicted that EM waves were governed by the same fundamental laws that light obeys. This was later confirmed by Hertz in 1887 who proved the photoelectric effect in electrons at the molecular level (Sadiku, 2007).

In order for an elecromagnetic wave to propagate it must have both an electric field component, and magnetic field component. In an EM wave, the electric and magnetic field components are orthogonal to each other, and orthogonal to the direction of propagation. Thus, EMI can only be caused by a time-varying field. A wave's field components and direction of propagation are related to each other by:

$$\hat{\mathbf{r}}_E \times \hat{\mathbf{r}}_H = \hat{\mathbf{r}}_\mathrm{P} \tag{2.40}$$

Where $\hat{\mathbf{r}}_E$, $\hat{\mathbf{r}}_H$, and $\hat{\mathbf{r}}_\mathrm{P}$ are unit vectors in the direction of the electric field, magnetic field, and direction of propagation, respectively. Figure 2.3 shows an example of a propagating wave.

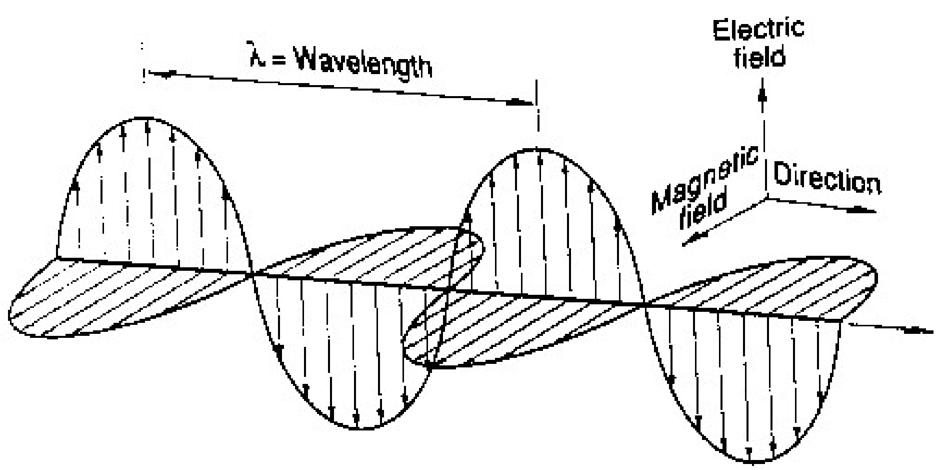


Figure 2.3. Propagating EM Wave (Mount Holyoke College, 2009).

A wave is a function of both time and space. Using the *Lorentz condition for potentials* the coupling between the electric and magnetic potentials can be solved.

$$\nabla \cdot \mathbf{A} = -\mu\varepsilon \frac{\partial V}{\partial t} \tag{2.41}$$

Using the Lorentz condition, and applying it to Maxwells equations, the equations for the electric and magnetic components of an EM wave can be determined.

$$\nabla^2 V - \mu\varepsilon \frac{\partial^2 V}{\partial t^2} = -\frac{\rho_v}{\varepsilon} \qquad \text{Elecctric Wave Equation} \tag{2.42}$$

$$\nabla^2 \mathbf{A} - \mu\varepsilon \frac{\partial^2 \mathbf{A}}{\partial t^2} = -\mu\mathbf{J} \qquad \text{Magnetic Wave Equation} \tag{2.43}$$

These equations are second order partial differential equations and will be revisited after disussing additional concepts relating to waves.

For an EM wave, electric and magnetic fields are proportional to each other. The ratio between them depends on the material surrounding the wave. The relative permittivity and permeability of the material determines its intrinsic impedance, for a lossless dielectric this can be expressed as:

$$\eta = \sqrt{\frac{\mu}{\varepsilon}} = \frac{|\,\mathbf{E}\,|}{|\,\mathbf{H}\,|} \quad (\Omega) \tag{2.44}$$

Substituting equations 2.4 and 2.15 into equation 2.44, the impedance for free space can be found. Free space has an impedance of 377**Ω**.

A wave has characteristics associated with it such as the velocity of propagation (v_p), angular frequency (ω), attenuation factor (α), and phase constant (β). These characteristics will all impact the wave as it travels in time. A wave will be examined that is traveling in free space. A perfect dielectric is classified as a material that has no conductivity or losses. Free space is lossless with no attenuation therefore, $\alpha = 0$. The angular frequency (ω) and phase constant (β) are given by:

$$\omega = 2\pi f \tag{2.45}$$

$$\beta = \omega\sqrt{\mu_0 \varepsilon_0} = \frac{\omega}{c} \tag{2.46}$$

Where $c \approx 3\text{x}10^8$ m/s and is the speed of light in a vacuum. For a forward traveling wave in free space, the electric and magnetic field are given by equations 2.47 and 2.48. These equations are solutions to the differential wave equations discussed previously (equations 2.42 and 2.43), and assume a sinusoidal time dependence of $e^{j\omega t}$ (Sadiku, 2007).

$$\mathbf{E} = E_0 \cos(\omega t - \beta z)\hat{\mathbf{r}}_x \tag{2.47}$$

$$\frac{\mathbf{H}}{\eta} = H_0 \cos(\omega t - \beta z)\hat{\mathbf{r}}_y \tag{2.48}$$

The velocity of propagation of a wave in free space is equal to the speed of light.

2.1.5. Skin Depth

The skin depth of a material determines how deep an EM wave is able to penetrate the medium. The skin depth is important to EMI because it will have an impact on the type of shielding that must be used. The depth the EM wave is able to penetrate is a function of the frequency and of the wave incident on the material, and the conductivity of the material. The skin depth is given by equation 2.49.

$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}} = \frac{1}{\alpha} \tag{2.49}$$

For copper, such as that used on a printed circuit board, the skin depth is shown as a function of frequency. Table 2.1 summarizes skin depth as a function of frequency.

Table 2.1.
Skin Depth in Copper (Sadiku, 2007).

Frequency (Hz)	10	60	100	500	10^4	10^8	10^{10}
Skin Depth (mm)	20.8	8.6	6.6	2.99	0.66	6.6×10^{-3}	6.6×10^{-4}

With a basic understanding of electromagnetics, it is understood how charge distributions are the source of electric fields, and current distributions are the source of magnetic fields. These fields both have an associated flux density based on their intensity. The electric and magnetic fields are able to couple to nearby devices and induce a voltage or current, respectively. The electric and magnetic fields are also able to couple to each other and radiate. This radiation is in the form of an EM wave and obeys the same laws as light. EM waves are quantized and can be used to transport energy or information. Now that EMI theory has been discussed EMC definitions will be introduced.

2.2. EMC Definitions

There are a wide range of terms commonly used when discussing electromagnetic interference (EMI), and electromagnetic compatibility (EMC). This section will expand on some of those terms, and introduce new definitions which will be used in later sections of this work.

As discussed before, there are two types of EMI emission types. These types are conducted and radiated. Conducted EMI is energy which has coupled to a power or signal bus, and is propagating through the system. Radiated EMI is interference which is no longer confined to a wire, but is an electromagnetic wave propagating away from the source. The source of EMI is wherever the interference is generated. A receptor for EMI is the device being affected by the interference. The amount the device is disturbed by unwanted EMI is referred to as susceptibility. Reducing the device susceptibility to EMI is also referred to as EMI hardening.

There are two different domains which can be used to analyze EMI. These domains are commonly referred to as the time and frequency domain. In the time domain, signals are analyzed as they change with time. In the frequency domain, signals are analyzed with respect to a specified band of frequencies. A tool commonly used to analyze signals in the time-domain is an oscilloscope, and in the frequency domain is a spectrum analyzer.

EMI energy can be both narrowband or broadband in nature. A signal that is narrowband is typically a repetitive signal, or pulse train. The signal can be a single frequency, a continuous wave (CW), or spectrally confined which makes it narrowband. Due to the repetitiveness of the signal, it is easy to make amplitude measurements. A signal that is broadband is typically a single pulse containing a span of multiple frequencies. This single pulse is typically very fast in the time domain. An example of broadband interference is from a spark-gap or lightning. Broadband signals can be difficult to measure since they are non-repetitive, and very fast. Signals which are slow or repetivtive in the time domain tend to be narrow in the frequency domain, and signals which are fast in the time domain tend to be broad in the frequency domain.

Due to the wide range of frequencies signals can exist at for EMC, most measurements are made using a logarithmic scale. For broadband or narrowband EMI signals are measured differently and therefore have different units. Narrowband EMI signals and their respective units are expressed in Table 2.2.

Table 2.2.
Narrowband EMI Signals (Mardiguian, 2001).

EMI Signal	*Units*
Voltage	V, dB 1V Reference (dBV) or dB 1µV Reference (dBµV)
Current	A, dB 1A Reference (dBA) or dB 1µA Reference (dBµA)
Power	W, mW or dB 1mW Reference (dBm)
E Field	V/m, µV/m or dBµV/m
H Field	A/m, µA/m or dBµA/m
Radiated Power Density	W/m^2, mW/cm^2, or dBm/cm^2

For broadband EMI, the received EMI is normalized to a unity bandwith. Broadband EMI signals and their respective units are expressed in Table 2.3.

Table 2.3.
Broadband EMI Signals (Mardiguian, 2001).

EMI Signal	*Units*
Voltage	µV/kHz, µV/MHz or dBµV/MHz
Current	µA/kHz, µA/MHz or dBµA/MHz
E Field	µV/m/kHz, µV/m/MHz or dBµV/m/MHz
H Field	µA/m/kHz, µA/m/MHz or dBµA/m/MHz

Coupling paths are a key issue in examining EMI problems. A coupling path is any means of transferring energy from the source to the receptor. Figure 2.4 shows common coupling paths for EMI.

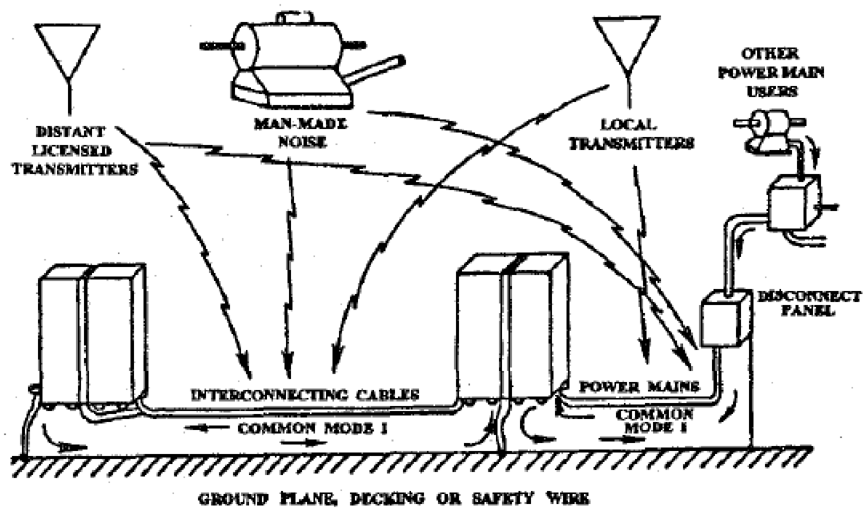


Figure 2.4. EMI Coupling Paths (Blanchard, & Sorak, 1994).

For any transfer of energy from the source to the receptor there is a coupling path. Common coupling paths are radiated fields to cabling and cross-talk. Identifying this coupling path and preventing this energy transfer is the first step in preventing EMI issues. Multiple coupling paths may exist in a system which should all be identified and isolated in order to increase EMI hardening.

Government standards are in place to limit the amount of EMI emitted from a device. The American standard is set by the Federal Communications Commission (FCC), and the European standard is set by the International Special Committee on Radio Interference (CISPR). These standards will be looked at more in depth in the following section, along with taking accurate EMI measurements in order to determine if a product is within compliance.

2.3. EMI Measurement Requirements and Techniques

Any product on the market which uses electronic circuitry must pass EMC requirements. With the ever increasing amount of electronic devices consumers use in their everyday lives, maintaining limits on the amount of interference each device can emit, while also receiving and still functioning properly, is important. These requirements are in place to protect devices from unwanted noise, such as noisy analog TV images, to malfunctioning completely such as dropped wireless communications. This section introduces EMI standards, EMI testing equipment, and techniques used to measure devices for EMC.

2.3.1. Regulatory Agency Requirements

When a product is tested for EMC a few preliminary questions must be answered. These questions include:

1. Where will the product be sold?
2. What is the classification of the product?
    - Industrial
    - Scientific
    - Medical
    - Information Technology Equipment (ITE)
    - Automotive
    - Communications
3. Where will the product be used?
    - Home
    - Commercial
    - Industry

The answers to these questions will define which standard the product must comply. Using the answers to the previous questions, Table 2.4 and Table 2.5 list the testing requirements the product must meet.

Table 2.4.
Agency Requirements for Emissions (Agilent Technologies, 2000).

CISPR	*FCC*	*EN's*	*Description*
11	Part 18	EN 55011	Industrial, scientific and medical
12	(SAE)		automotives
13	Part 15	EN 55013	Broadcast receivers
14		EN 55014	Household appliances / tools
15		EN 55015	Fluorescent lights / luminaries
16			Measurement apparatus / methods
22	Part 15	EN55022	Information technology equipment
		EN50081-1,2	Generic emissions standards

Table 2.5.
European Requirements (Agilent Technologies, 2000).

Equipment type	*Emissions*
Generic equipment	EN 50081-1
Residential	
Light industrial	
Industrial	EN 50081-2
Information technology equipment (ITE)	EN 55022
Industrial, scientific, medical products (ISM)	EN 55011

From Table 2.4 it can be seen that CISPR and FCC are the two major committees which set standards for EMC compliance.

The FCC and CISPR have strict regulations in order to maintain a firm grasp on how much EMI a device can emit. Most of this information is outlined in FCC Part-15 Subpart-B and CISPR 22, which define radiated emission limits for unintentional radiators. Electronic devices are grouped into different classes. Class A products include devices which are sold to schools and other commercial institutions and Class B products which are sold in residential areas (Federal Communications Commission, 2005).

Table 2.6, Table 2.7, and Table 2.8 provide the minimum requirements which must be passed in order to become certified by the FCC or CISPR.

Table 2.6.
FCC: Class A Products at a distance of 10m (Federal Communications Commission, 2005).

Frequency of Emission (MHz)	Maximum Field Strength (dBµV/m)
30 - 88	90
88 - 216	150
216 - 960	210
Above 960	300

Table 2.7.
FCC - All devices excluding Class A at a distance of 3m (Federal Communications Commission, 2005).

Frequency of Emission (MHz)	Maximum Field Strength (dBµV/m)
30 - 88	100
88 - 216	150
216 - 960	200
Above 960	500

Table 2.8.
CISPR – Quasi-Peak measurements at 10m (CISPR 22, 1997).

Frequency of Emission	Maximum Field Strength (dBµV/m)	
(MHz)	Class A	Class B
30 - 230	40	30
230 - 1000	47	37

2.3.2. EMI Measurement Process

All radiated EMI measurements must be performed by an FCC certified company. This is to insure the equipment is capable of taking accurate measurements and is calibrated correctly. The tests are performed in a large anechoic or semi-anechoic chamber so ambient RF energy is shielded. A high performance ultra-wideband bilogrithmic antenna is commonly used for EMI testing. The bilogrithmic antenna (Figure 2.5) is a biconical antenna and log-periodic antenna combined together. The combined antenna exhibits a linearly polarized pattern, and is very accurate when detecting electric fields. Combining these two antennas also decreases the time it takes to perform a test.

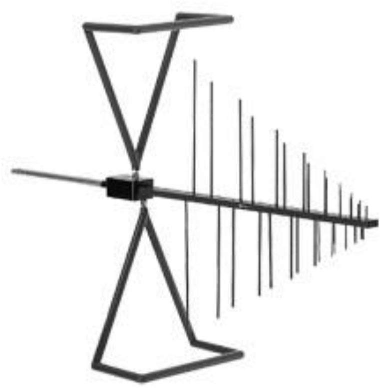


Figure 2.5. Bilogrithmic Antenna (Teseq Worldwide, 2009).

When performing an EMI test, the equipment under test (EUT) is placed in the anechoic chamber on a rotating carousel. The EUT is then powered up and the antenna moved 3m or 10m away depending on what the standard calls for. The exact test setup can be seen in Figure 2.6.

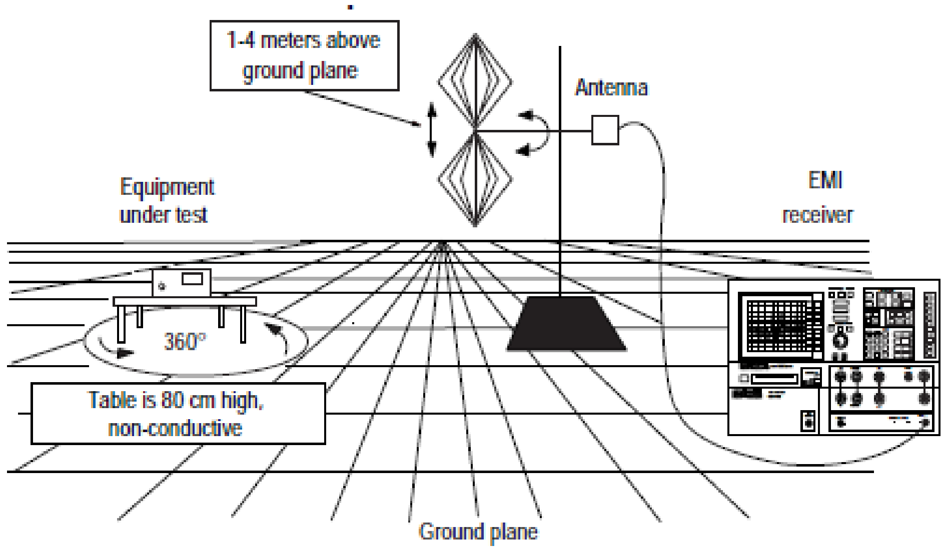


Figure 2.6. Radiated EMI Test Setup (Agilent Technologies, 2000).

The bilogrithmic antenna is mounted on a mast which has two motors attached to it. One motor moves the antenna vertically, and another rotates the antenna's azimuth 90 degrees. A series of measurements are made with the antenna at different heights, while the carousel that the EUT sits on, rotates a full 360 degrees. The antenna's azimuth is then rotated 90 degrees and the test repeated.

The antenna measurements are captured with an EMI receiver. The EMI receiver is very similar to a spectrum analyzer, however is specifically tailored for EMI tests. Depending on the standard, the EMI receiver will either perform a peak, quasi-peak or average measurement on each detected signal. For signals below 1GHz, both the FCC and CISPR requirements call for a quasi-peak measurement. For signals above 1GHz the FCC requires an average measurement, and CISPR has no regulations. The quasi-peak measurement system is based off of the signal repetition rate. Quasi-peak detectors have a much faster charge rate than discharge. As a repeated signal is detected the EMI receiver charges faster than discharges which results in a higher output voltage. An average measurement is done simply by taking multiple measurements.

The recorded measurements are plotted over the measured frequency span. This frequency span is plotted from 30MHz to 1GHz for both the FCC and CISPR standards. The FCC has provisions which allow for EMI measurements up to 5GHz. Measurements to 5GHz are only required if the product has an oscillator at 108MHz or faster. When the EMI receiver plots the recorded measurements it also plots a line for the allowed FCC limits. From the plot, it is obvious if any frequencies are exceeding the allowed radiated emissions. Appropriate steps should then be taken to track down the problem, and reduce the unwanted emissions.

Other standards exist which specify limits on conducted EMI and device susceptibility to EMI. Specifically, the conducted EMI standard states there is a limit to the amount of conducted EMI allowed to travel from a device back onto the power grid. Special tests are also performed which test devices EMI susceptibility by placing it in a very noisy environment, and monitoring it for malfunctions. In order to become EMC compliant a device must also pass these other tests. The compliance standards are beyond the scope of this work. The primary focus of this work is on unintentional radiated emissions.

2.3.3. Locating Unintentional EMI

Tracking down unintentional radiators is a difficult task. Once the plot of the electric field strength vs. frequency is created by the testing facility, it should be clear which frequency is causing issues. Typical testing areas are not the most ideal places for troubleshooting EMI issues, they are typically busy work areas and don’t have the equipment to track down radiators to the circuit board level. A solution is needed to track down EMI problems in a lab environment. This solution comes in the form of EMI sniffer probes.

Using EMI sniffer probes along with a spectrum analyzer, real time EMI measurements can be made from a lab environment which allows for changes can be made on the fly. Using the plot from the testing center, problem frequencies can be tracked down to the location where the radiation is largest. There is no correlation between measurements made by the test center and the sniffer probes; however, the amount of reduction in radiated emissions can be measured. There will be a correlation between the amount of reduction in emissions measured by the sniffer probes, and measured by the testing facility.

There is more than one type of sniffer probe available. Both electric and magnetic field sniffer probes exist, along with probes designed to measure EMI directly by touching the circuit board or wire. Sniffer probes are an excellent and cost effective way to troubleshoot EMI issues. Now that unwanted EMI has been located, common problems which result in unintentional radiated emissions, and their solutions, will be introduced in the following sections.

2.4. Common EMI Problems

There are multiple EMI problems which can cause an electronic device to act as a source of interference. Fast switching currents and voltage slew rates, *di/dt* and *dv/dt*, have major impacts on the amount of EMI a device produces. This section will provide an overview of some of the most common problems which contribute to EMI, and later work will discuss solutions to these problems.

As discussed previously, there are two types of EMI that need to be evaluated. These types are conducted and radiated. Conducted EMI problems will be discussed first, before analyzing radiated EMI issues. Most conducted EMI problems reside with the power supply. As power supply component density, and switching frequency increase, so do the chances for EMI problems. Conducted EMI is high frequency noise picked up or generated on the power bus. This noise can come from the main power station, other devices connected to the electrical grid, a noisy source located next to power lines, or the power supply of the main device. There have been a few case studies on power lines located next to communication broadcasting antennas. The radiated energy from the antenna coupled onto the power lines, and was fed back into the station which damaged sensitive equipment. Not all scenarios may be this critical, but it is important to recognize conducted EMI is an important issue, and that devices must filter out incoming interference, while also blocking outgoing interference to the electrical grid. Conducted EMI can be also created by how the electronics fit in the enclosure of the device. Internal wiring has the potential to couple energy to each other. Wiring passing over sensitive devices such as crystal oscillators or particularly noisy sections on a PCB such as analog signals, high frequency digital signals, or RF signals may also couple energy onto a cable. This cable then has the potential to radiate the conducted EMI.

2.4.1. Cross-Talk

Cable to cable coupling, or noisy PCB element to cable, are two methods of internally generating conducted EMI problems. There are other types of coupling which primarily take place on a PCB that are due to mutual inductance, or mutual capacitance. This coupling is commonly referred to as cross-talk. Cross-talk is caused due to near-field coupling between signals located in very close proximity to each other. Capacitive cross-talk is the most common and occurs due to very fast voltage slew rates (*dv/dt*). The quickly changing voltage waveform on the first signal induces a voltage on the signal running parallel to it. The simple definition of capacitance is two metal plates next to each other. The actual amount of capacitance formed relies on geometry of the structure. Capacitance can be calculated using the following equation:

$$C = \frac{\varepsilon A}{d} \tag{2.50}$$

As can be seen in equation 2.50, the effective area (A) of the parallel traces, and the separation between (d) are critical to determine how much effective capacitive coupling (C) will occur. To accurately calculate the mutual capacitance between two parallel traces the fringing fields must be accounted for. Equation 2.50 cannot be used to calculate mutual capacitance; it was used to show how geometry will impact mutual capacitance. Figure 2.7 shows an example of capacitive cross-talk.

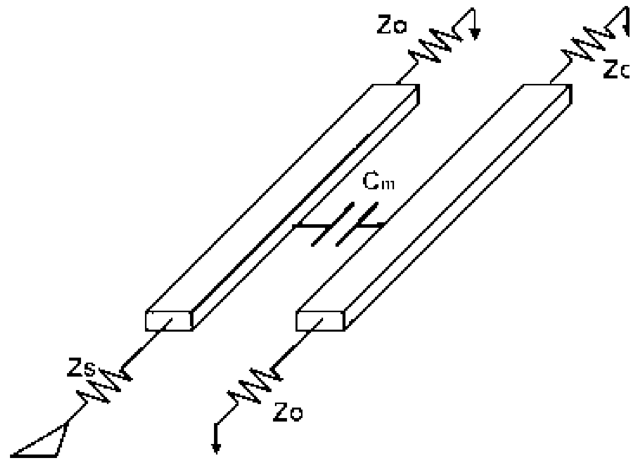


Figure 2.7. Capacitive Coupling Cross-talk (Intel Corporation, 2002).

The second and less common form of cross-talk is inductive. Inductive cross-talk occurs from mutual inductance and magnetic coupling. A pulse of current (*di/dt*) traveling down a signal path will produce a magnetic field surrounding it. This magnetic field couples to the parallel signal path which induces a current in the opposite direction. The current traveling in the opposite direction can be explained from Lenz's law, and is due to the opposition in change of magnetic flux (Intel Corporation, 2002). An example of inductive cross-talk can be seen in Figure 2.8. The frequency and amplitude of the voltage will play an important role determining if cross-talk will occur. Higher frequency RF signals and large voltage or current transitions will provide better chances for cross-talk.

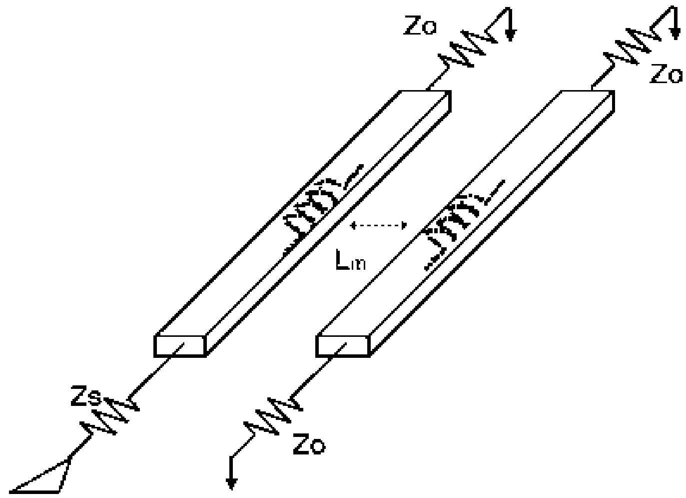
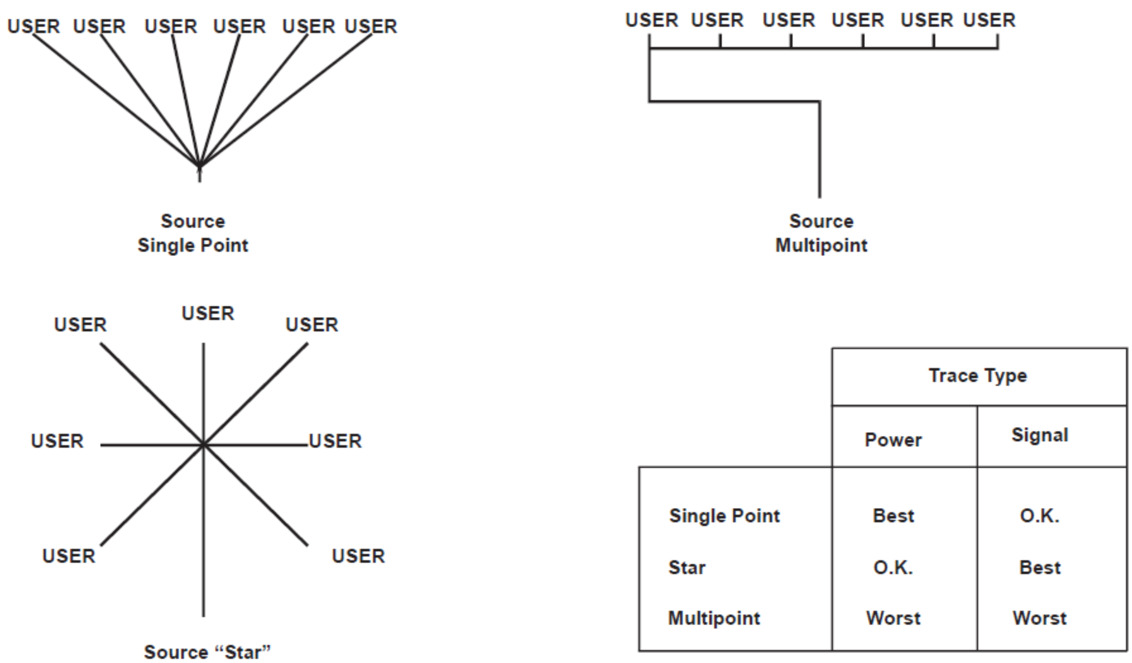


Figure 2.8. Inductive Coupling Cross-talk (Intel Corporation, 2002).

Cross-talk involves capacitive or inductive coupling which induces interference on a nearby signal. Other types of conducted EMI exist and are formed by ground loops. Ground loops can be very problematic. In audio systems ground loops can cause 50/60Hz hum, and in video systems noisy images. Ground loops are a result of multiple return paths for current which all have different impedances.

2.4.2. Ground Loops

In power systems, ground is commonly tied to earth, however in digital electronics, ground is normally a reference potential. In digital electronics, ground reference is typically, but not always, tied to 0V. In a system with multiple return paths each ground may be at a slightly different potential. This voltage potential can be due to many reasons, but commonly is due to non-perfect conductors. As of today, a perfect conductor does not exist. This means that anything which conducts electricity has some non-ideal characteristics. One of these characteristics being resistance per unit length. Different lengths of wire will have different amounts of resistance. If a ground loop is formed, the resistance of the cable will drop a small voltage. The potential different between grounds can cause many problems, some being noise, and others can prevent the system from operating normally. Different ground configurations can be used depending on the system which perform better than others in terms of reducing noise. Some of these ground configurations are shown in Figure 9.



	Trace Type	
	Power	Signal
Single Point	Best	O.K.
Star	O.K.	Best
Multipoint	Worst	Worst

Figure 9. Grounding Configurations (Texas Instruments, 1999).

Ground loop noise can be classified into separate categories. These are common-mode inferference, and differential-mode interference. Common-mode interference is the most common source of noise on PCBs with fast switching digital logic (Texas Instruments, 1999). Common-mode impedance is caused by impedance that is tied to both the source and the load. This impedance has a voltage potential across it (Figure 2.10), which causes the current flowing to the load to travel in the same direction on both the source and return path. This impedance is generally due to different paths to the same common ground.

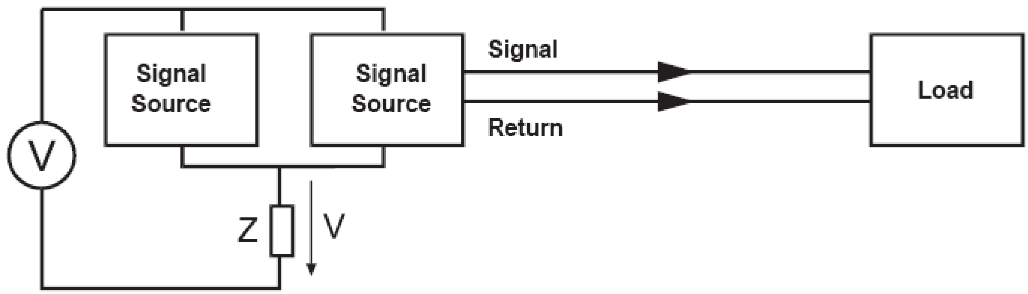


Figure 2.10. Common-Mode Interference (Texas Instruments, 1999).

Differential-mode interference is caused by the difference in voltage potential between the source and return trace. This difference in voltage potential is caused as current flows from the source through the load, and back to the source. Figure 2.11 shows an example of differential-mode noise. The load drops all of the voltage creating the difference in potential. Differential-mode noise is unavoidable and can only be minimized (Texas Instruments, 1994).

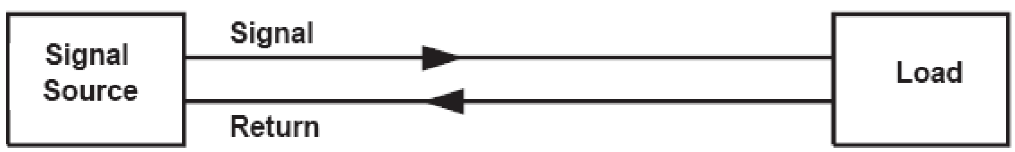


Figure 2.11. Differential-Mode Interference (Texas Instruments, 1999).

Conducted EMI is interference traveling down a signal path that has entered the system through the main power bus, coupled to the device via other signals in close proximity, or formed by ground loops. The conducted EMI is always looking for a return path to the source. When energy couples onto a nearby signal path, this return path often becomes very complex, or even non-existant. A very long return path often involves many loops or obstacles essentially forming an antenna capable of radiating or coupling noise into the system. If no return path exists the energy will radiate. Radiated emissions are typically due to antennas which are unknowingly formed due to cabling, or signal paths on a PCB.

2.4.3. Antenna Effects

Radiated EMI problems can be caused by currents with convoluted return paths, lengthy signal paths on a PCB, or conducted EMI with no return path. Radiated EMI can also be due to antennas formed by the devices internal or external cabling. As discussed earlier, every frequency has an associated wavelength. The wavelength can be calculated using the following equation:

$$\lambda = \frac{c}{f\sqrt{\varepsilon_r}} \qquad (2.51)$$

Where λ is the wavelength, c is the speed of light, ε_r is the relative permittivity, and f is the frequency. From equation 2.51 it can be seen that as the frequency increases, the wavelength decreases. As the wavelength becomes smaller, signal paths and electrical components begin to appear electrically large. When something appears electrically large when compared to wavelength, it is implying that multiple wavelengths are capable of fitting onto it. When multiple wavelengths fit on a signal path, the path can no longer be considered a node, where the voltage is constant across the entire path, but it must be considered a transmission line where voltage and current vary as a function of position. When designing antennas, the wavelength of the operating frequency is a critical variable. A monopole a very basic type of antenna which can be made from a single wire.

The wire needs to be long enough that λ/4 fits onto it. Figure 2.12 Shows an example of a monopole antenna.

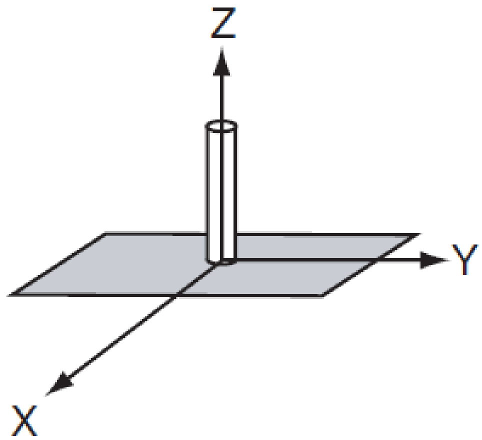


Figure 2.12. Monopole Antenna (Naval Air Systems Command, 1997).

A dipole antenna is a simple antenna formed when two monopole antennas are combined. Figure 2.13 shows an example of a dipole antenna.

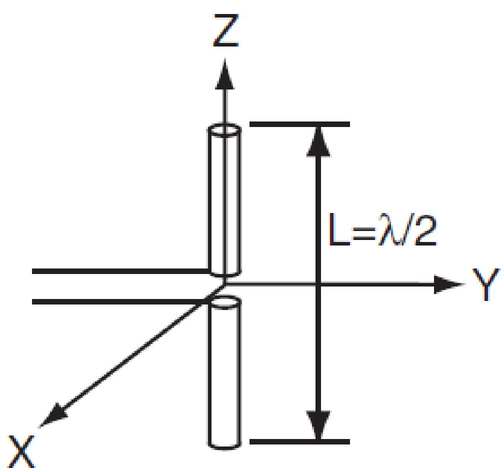


Figure 2.13. Dipole Antenna (Naval Air Systems Command, 1997).

The monopole and dipole antennas are excited by electric fields inducing a voltage on them. Antennas can also be excited by magnetic fields inducing currents. A loop antenna is a very basic antenna for coupling magnetic energy. The loop antenna can vary in length depending on the amount of gain, or directivity, that is required. Loop antenna lengths typically vary from a loop that is λ/4 to a complete wavelength in circumference.

When discussing antennas it is important to note that they are reciprocal devices. This simply means that any antenna which is used to transmit electromagnetic energy, will behave in the exact same manner when receiving electromagnetic energy. Basic examples of antennas were discussed because of how simple they can be, and the potential to unknowingly create them. A simple wire used which is 0.5m in length can also form a monopole antenna capable of radiating fairly well at 150MHz. This is shown as follows:

$$\frac{\lambda}{4} = 0.5\mathrm{m}$$

$$f_{rad} = \frac{c}{4(0.5)} = 150\mathrm{MHz} \qquad \text{Monopole Frequency} \tag{2.52}$$

A wire which is 0.5m is fairly long, but the principal applies to shorter wire or PCB interconnects as the frequency increases. As the length becomes shorter, the frequency at which the monopole radiates increases. Since the antenna is also a reciprocal device, the connection will also couple energy into the device at that frequency. Identifying potential radiators, and minimizing effective loop areas will reduce radiated EMI and reduce EMI susceptibility.

2.4.4. Spark Gaps

Another source of radiated EMI is from spark gaps. Spark gaps exist in nature, and are also used in many electrical devices. In nature a spark gap is caused by lightning, or electro-static discharge (ESD). Common uses of man-made spark gaps today are in automobiles such as spark plugs, and shunting devices for high voltage surges. Many times the material in in the gap is air. When a large voltage potential exists, the air in the gap becomes ionized. The ionized air has a lower resistance and a current is able to travel through it. An electrical arc will form in the gap area, and will last until the voltage potential lowers to a level that is unable to sustain it.

The spark gap creates an electrical pulse of energy that is very fast in the time domain. As discussed previously, any event that is very fast in the time domain tends to be broad in the frequency domain. Spark gaps emit electromagnetic radiation across the entire frequency spectrum. Engineers recognized this in the early days of wireless communication, and spark gaps were used to send Morse code. Today more sophisticated methods are used to send data wirelessly, but spark gaps are still used in everyday devices. Devices which operate near man made spark gaps must be protected accordingly. Static electricity and lightening are known for destroying many electrical devices including computers. Some components are designed to be ESD resistant, and are rated to withstand voltage discharges up to several kilovolts. The voltage potential with lightning is so large that devices have little chance of surviving if conducted EMI couples to the power grid. Protective devices have been designed to shunt this excess energy away from the device, these devices will be further discussed in following sections.

There are techniques which can be implemented to reduce both conducted and radiated EMI emissions. Most, if not all, devices must implement some type of EMI hardening technique in order to meet EMC requirements.

2.4.5. Impedance Mismatch

A properly designed system will have matched impedances from the source, to the transmission line, to the load. This insures maximum power transfer, and minimizes reflections. Reflections on the transmission line maximize current, which also increase radiated emissions. The transmission line can be a cable attached to the system, or a copper trace on a printed circuit board. Unmatched impedances will cause digital signals to overshoot and ring and result in additional radiated emissions. Properly matching impedances will reduce the amount of radiated emissions from the device.

There are multiple ways of impedance matching depending on the design requirements. When dealing with RF systems maximum power transfer is often critical. Impedance matching methods used in RF systems can be L-networks, Pi-networks, T-networks, or distributed parameter systems. These matching networks involve purely reactive elements for matching, and don’t dissipate any energy. Impedance matching for digital designs, as relating to EMI, are less concerned with maximum power transfer and more concerned with signal overshoot, undershoot, and ringing.

When matching impedances, the source and load impedances must be known. For RF systems the characteristic impedance is typically 50Ω, and in video devices it is 75Ω. When discussing digital clock drivers and line drivers, the input impedance can vary between 10kΩ and 100kΩ.

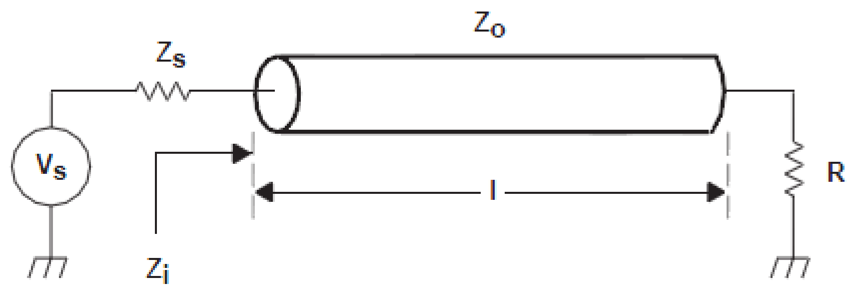


Figure 2.14. Transmission Line Model (Texas Instruments, 1994).

From the transmission line model (Figure 2.14), Z_s is the source impedance, Z_o is the transmission line impedance, l is the length of the transmission line, R is the load impedance, and Z_i is the input impedance as seen at the input of the transmission line. For a properly matched system the source impedance, transmission line impedance, and load impedance should all be equal. When it is impossible to have identical impedances, matching must occur. Unmatched impedances will cause reflections on the transmission line; these reflections maximize current on the transmission line and cause ringing exacerbating radiated EMI.

To determine the amount of ringing which will take place on the transmission line the reflection coefficient can be calculated. The reflection coefficient determines how much of the signal will be reflected due to an impedance mismatch. The reflection coefficient is calculated as:

$$\Gamma = \frac{Z_L - Z_o}{Z_L + Z_o} \tag{2.53}$$

In a perfectly matched system the reflection coefficient will be 0, with no reflections. The reflection coefficient can vary between +1, where the load is open, to -1 where the load is shorted.

There are multiple types of transmission lines ranging from coax cables to a printed microstrip or stripline. A coax cable has impedance which is determined by the ratio of the inner and outer diameters of the cable. Common impedances for coax cables are 50Ω and 75Ω. A microstrip is a transmission line printed on a circuit board and can be theoretically designed at any impedance. The microstrip is a printed trace above a ground plane. Figure 2.15 shows a microstrip transmission line.

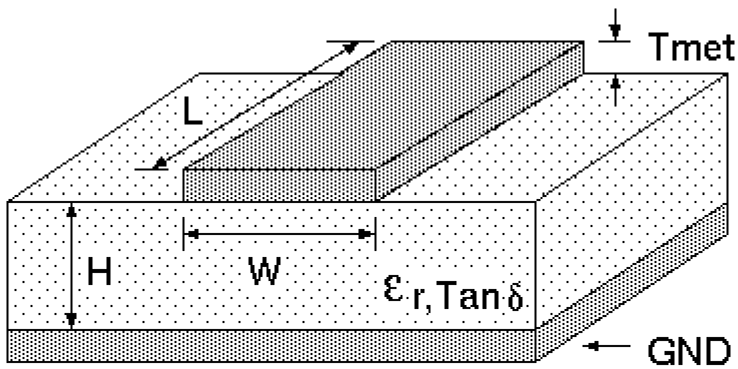


Figure 2.15. Microstrip Transmission Line (McMahill, 2003).

Impedance matching designs can be used either before or after the transmission line. For a transmission line which will be used in bidirectional applications, matching should take place after the transmission line. For single direction applications matching should take place before the transmission line (Texas Instruments, 1994). Matching designs will be further discussed in the EMI solutions section of this work.

2.4.6. <u>Harmonics</u>

The last EMI problem discussed in this chapter has to do with frequency harmonics and cross-products. Fourier theory states that any periodic signal can be broken down into a series of sine and cosine signals. These sine and cosine signals are multiple integers of the periodic signal, and are called harmonics. The superposition of all the harmonic content defines the shape of the waveform. For example, a square wave used in digital electronics is a non-sinusoidal waveform. The square wave consists of only odd integer harmonics (i.e. 3^{rd}, 5^{th}, 7^{th}). A saw-tooth waveform is also non-sinusoidal, but consists of both odd and even integer harmonics (i.e. 2^{nd}, 3^{rd}, 4^{th}). This is important when discussing EMI because often times radiated emissions are not caused by only the fundamental frequency, but from the harmonic content as well.

2.5. EMI Solutions

In order for a device to pass EMC requirements, it must satisfy the requirements set by the American FCC, or European CISPR standards. In the past many designers would simply ignore EMI issues until the final stage of development. Before the product was to hit the market it would be tested for EMC. Until the product passed, it could not be bought or sold. Many times this would result in a product delay to market until engineers could fix the problem, and a substantial increase in cost. Figure 2.16 shows the cost needed to correct EMI problems throughout the product development cycle.

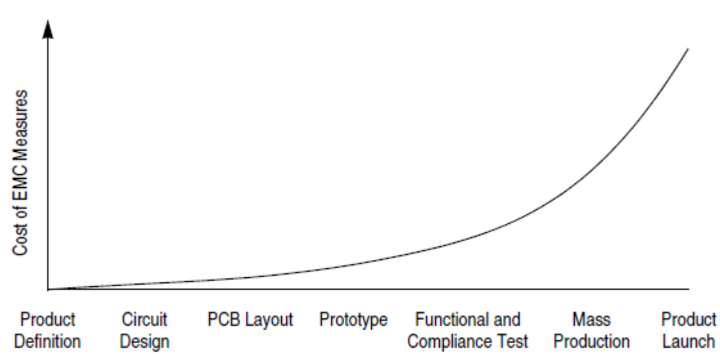


Figure 2.16. Cost of EMI Corrections (Freescale Semiconductor, 2005).

In some cases simple EMI hardening techniques could be used to reduce the emissions, other times the product would have to be completely redesigned. As a solution, the EMC expert should be actively involved in the product design (Nageswara Rao, Venkata Ramana, Krishnamurthy, & Srinivas, 1995). Figure 2.17 shows a product development cycle with an EMC engineer involved.

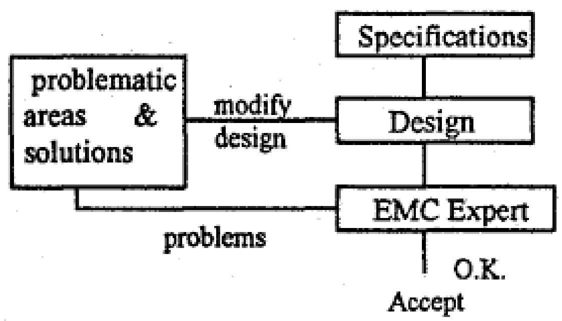


Figure 2.17. Product Development Cycle
(Nageswara Rao, Venkata Ramana, Krishnamurthy, & Srinivas, 1995).

This section will discuss techniques to reduce unwanted EMI emissions, and hardening techniques that can be used to decrease device susceptibility.

When discussing EMI solutions, the most critical design aspects will be analyzed first. These solutions are discussed first because they are not easily changed later in the product design cycle. These solutions include proper circuit design along with board layout and zoning techniques which should be implemented first. Later solutions will include suggestions which can be made quickly such as ferrite material, enclosure shielding and power line filters. The first solution will discuss capacitive decoupling.

2.5.1. Capacitive Decoupling

One method to minimize the effective loop area of a power bus is to use decoupling capacitors. A properly placed decoupling capacitor should be located as close to the power pin on the associated device as possible. This decoupling capacitor looks like the power source to the device and therefore minimizes the loop area. The decoupling capacitor also shunts out common-mode noise which may have coupled onto the bus. Figure 2.18 shows an example of using a decoupling capacitor to minimize loop currents.

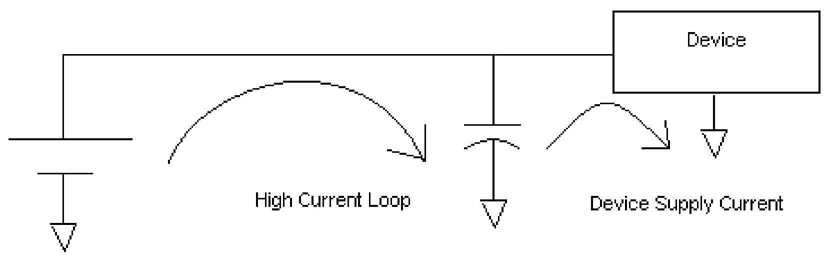


Figure 2.18. Decoupling Capacitor to Minimize Loop Current.

2.5.2. Circuit Board Layout

Proper board zoning is critical, reducing EMI starts at the board level. Preventative measures taken with proper layout could reduce EMI problems later on. Proper board zoning will minimize unwanted coupling and reduce the amount of later solutions needed such as shielding the enclosure and expensive shielded cables. A properly zoned circuit board should have the digital, analog, RF and other sensitive circuits separated, so that each section has its own ground (Texas Instruments, 1999). Placement of high frequency components like switching regulators and microcontrollers should be located near one another and slower frequency circuitry such as audio should be placed farther away. Figure 2.19 shows an example of a properly zoned circuit board.

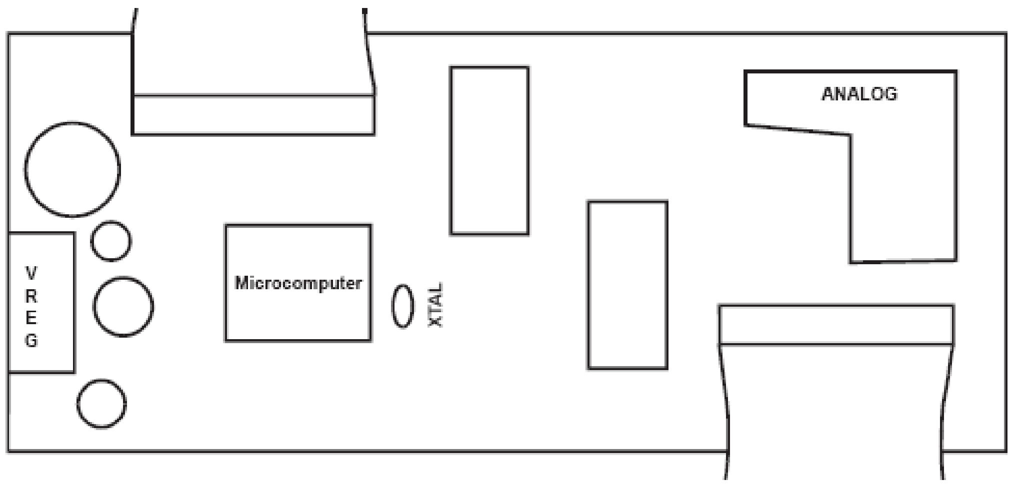


Figure 2.19. Proper Board Zoning Layout (Texas Instruments, 1999).

It is critical to keep analog traces away from switching digital lines and clocks so that there is no coupling between the two. This coupling, as mentioned before, can create a large return path loop for the current causing it to radiate. It is equally important to keep RF traces away from others to prevent crosstalk.

A second method to reduce effective loop areas is to use a multi-layer circuit board. A multi-layer circuit board consisting of 4 of more copper layers has been shown to reduce radiated EMI over a standard two-layer circuit board. In a two-layer circuit board all effective loop areas are larger. This includes separation between the VCC and ground traces, decoupling capacitor loop, and signal trace to ground loops. The loop areas are effectively larger due to a few reasons. The signal and power paths must be routed on the same layer. In a two-layer board these signals will have to be routed around one another, and the path length will be increased. Figure 2.20 shows an example layout with a two layer board where signals may have to be routed around one another.

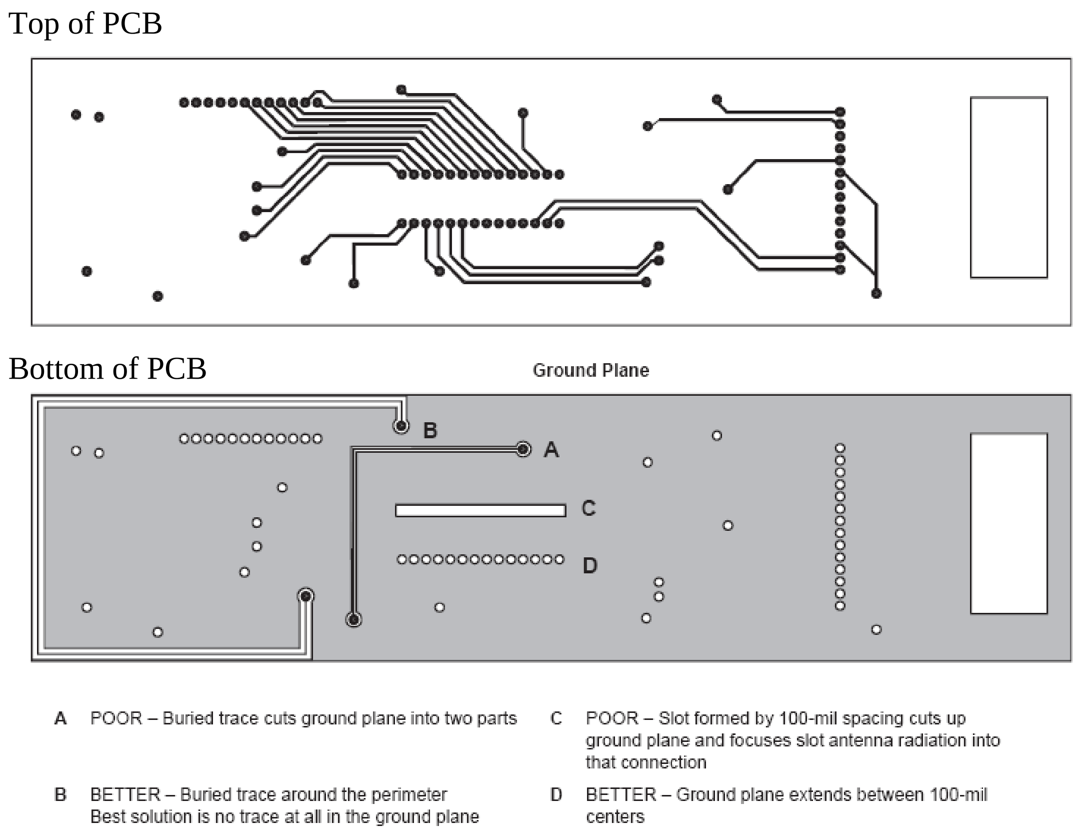


Figure 2.20. Two-Layer PCB with “Broken” Ground (Texas Instruments, 1999).

Current prefers to return using the same path from which it was sourced. This means it is ideal to have the current return path directly under the signal or power source path. In a two-layer board this is nearly impossible to do. The solution is a multi-layer circuit board. The multi-layer circuit board is typically configured as shown in Figure 2.21.

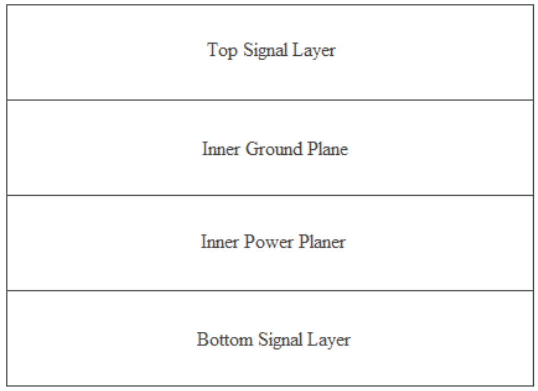


Figure 2.21. Multi-Layer Circuit Board Layout.

Using the multi-layer circuit board configuration shown in Figure 2.21, the ground and power planes reside in between the signal and layers. This provides a few benefits over two-layer boards. The first benefit is that all loop areas are effectively minimized. By having a dedicated power and ground plane the effects of the broken ground plane seen in Figure 2.20, are minimized or completely eliminated. Another benefit of a multi-layer board over a two-layer is that the distance between copper layers is less. Typical circuit boards are approximately 1.6mm thick. A two-layer circuit board has epoxy filling the gap between layers. A multi-layer circuit board has additional layers, and therefore the separation between layers is less. By having a solid ground plane under the signal layer it also provides some degree of shielding by terminating fringing fields.

As the designer or layout engineer creates the PCB, there is always the option to use an auto-router to complete the board. For RF and noise related issues, this isn't always the best method. The auto-router has no concept of EMI and therefore cannot take any preventative measures. The software has no way of knowing if one signal over the other will be more susceptible to noise, or if one trace carries an analog signal while the next carries a digital clock. The software also has no way of recognizing that a component is actually an oscillator and that routing around that object should be avoided. Critical traces should be routed first and keep-out zones defined before using the auto-router. A prioritized list needs to be followed when laying out a PCB. If the auto-router is used the PCB should be carefully examined for EMC issues.

Proper PCB layout techniques are important for reducing EMI emissions, however proper circuit design also plays an important role. Cross-talk between parallel signal paths, both capacitive and inductive, can induce common-mode noise throughout the system. Differential mode noise may also exist due to lengthy return paths. This coupling must be accounted for and attenuated properly. Impedance matching can also be implemented to reduce overshoot and ringing. Design considerations should be implemented assuming worst case scenarios. It is always easier to design EMI solutions into the layout, and remove what is not needed, rather than return at a later date and fix problems or add components.

2.5.3. Ferrites

During the schematic layout, ferrites should be used in the system to reduce common mode and differential mode interference. There are multiple types of ferrites for use in different applications. Ferrites can be used in RF designs, such as couplers or circulators, where low loss and high efficiency is desired. There are also ferrites for use in EMI. These ferrites are expected to have high loss and low efficiency and their sole purpose is to attenuate interference. Ferrites for EMI are made of a lossy material which has a high permeability. As discussed previously, the high permeability allows for magnetic field lines to travel easily through the material. Typically ferrites are thought of as inductors, however they act more like a lossy transformer. The impedance of a ferrite is given as:

$$Z = \sqrt{R^2 + L^2\omega^2} \tag{2.54}$$

Ferrites can be used in different configurations to attenuate common mode and differential mode noise. By placing a ferrite on both the signal and return path both common mode and differential mode noise is attenuated, where as by placing individual ferrites on both the signal and return only differential mode is attenuated (Mardiguian, 2001). Figure 2.22 shows an example of ferrites used in both configurations.

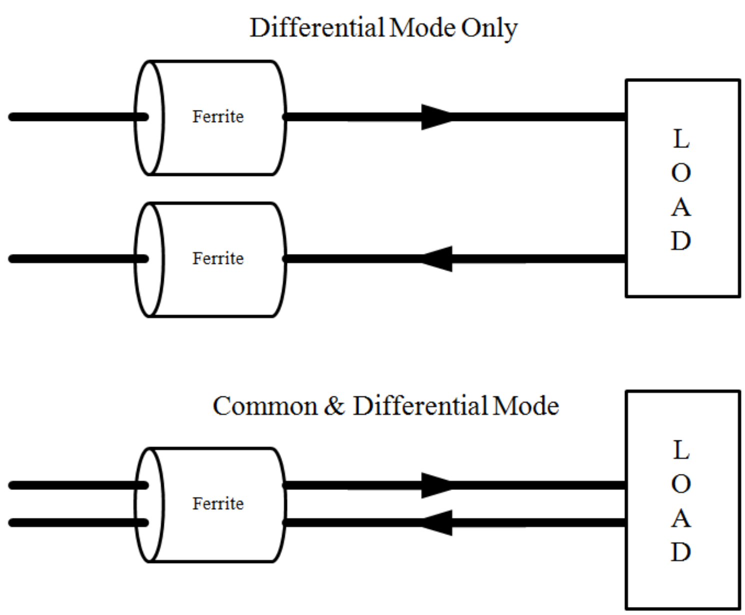


Figure 2.22. Ferrite configuration for reducing CM and DM noise.

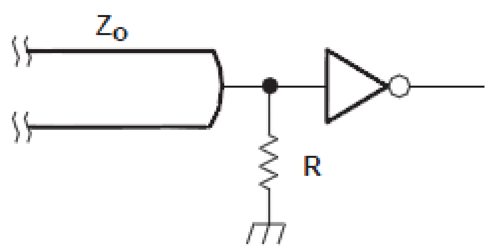
Careful consideration must be taken when using ferrites to reduce both common and differential mode noise at the same time. Any high frequency signal which is intentional can be disturbed by this configuration. Typically ferrites are used on the power bus feeding the device. By placing ferrites on the power bus any common mode or differential mode noise is attenuated before feeding the device, and there is no risk of disturbing any intentional high frequency signals. This solution plays an important role when it comes to proper board zoning.

As discussed previously, the PCB should be zoned properly to isolate the analog, digital and RF circuitry. Each section of the board should have its own individual power and ground. Each of the sectioned grounds should be tied to a single ground through a ferrite forming a star ground pattern. The power supply path to each section should also pass through a ferrite before feeding the decoupling capacitors for the device. Utilizing ferrite beads on the power and ground paths will attenuate any common mode or differential mode noise before supplying power to the device, and attenuate any noise generated by the device. Using ferrites in designs is an excellent way to reduce radiated EMI issues.

2.5.4. Impedance Matching

Additional circuit design considerations is implementing some form of impedance matching for high frequency digital switching busses. Impedance matching is important to prevent signal ringing. This ringing maximizes the current on a signal path, and contributes to EMI. If bidirectional data flow is needed, impedance matching should be done at the end of the transmission line. In contrast if the data is unidirectional impedance matching should be done before the transmission line. A few methods are presented in order to match a transmission line.

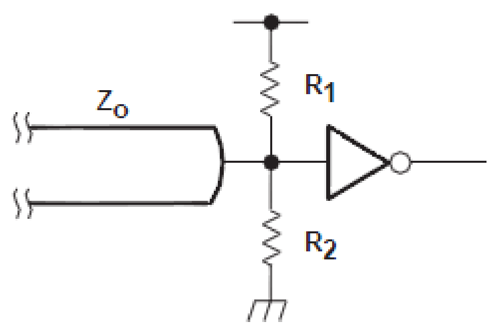
The first method presented uses a simple pull-down resistor (Figure 2.23). Without this pull-down resistor the transmission line is not terminated, and is connected to a buffer with a large input impedance. This impedance mismatch would cause reflections on the line and contribute to EMI. A simple pull-down resistor can be used to terminate the transmission line. In order to match the system the pull-down resistor should have the same characteristic impedance as the transmission line.



$$Z_o = R$$

Figure 2.23. Impedance matching with a pull-down resistor (Texas Instruments, 1994).

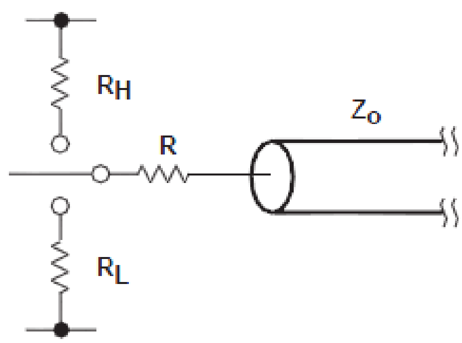
A second method can be used to reduce power consumption if necessary. The second method uses a Thevenin termination technique. This method also has an additional resistor, which if board space is a concern may be more difficult to implement. Figure 2.24 shows a Thevenin termination using a parallel network.



$$Z_o = \frac{R_1 \times R_2}{R_1 + R_2}$$

Figure 2.24. Impedance matching with parallel resistors (Texas Instruments, 1994).

The previous information on matching networks assumed that bidirectional data was needed, and therefore impedance matching was done after the transmission line. Both matching techniques are referred to as Thevenin terminations. For unidirectional data flow, a damping resistor is used on the input of the transmission line. Figure 2.25 shows a damping resistor matching network.



$$Z_o = \frac{R_H + R_L}{2} + R$$

Figure 2.25. Impedance matching with a damping resistor (Texas Instruments, 1994).

Not only is proper board zoning and circuit design critical, but the enclosure and peripherals around a device are also important. Internal wiring can couple to fields radiated by the circuit board and in return act as an even larger antenna, and contribute to conducted EMI. External wiring carrying common mode and differential mode noise can act as antennas and radiate. Use of shielding, and ferrites can reduce some of these additional problems.

2.5.5. Cabling and Shielding

As devices become smaller, the distance between signals are reduced on the circuit board, and with any cabling within the enclosure. This internal cabling is capable of coupling with fields and signals on the circuit board just like parallel signal paths. The internal cabling may then contribute to conducted EMI problems. This conducted EMI could be transmitted to other circuit boards in the product or exit the enclosure and run to other devices. Some of these problems can be fixed quickly after the board design by using shielding, ferrites, and filters.

For internal cabling, careful routing considerations must be made. Internal routing should avoid RF sections of circuit boards and other particularly noisy sections. As a general rule, no trace or wire should pass closer than 25.4mm around an oscillator (Texas Instruments, 1994). For cabling such a ribbon cables guard paths may be used to prevent coupling. Using guard paths simply means placing a ground between signals. For ribbon cables this means alternating signal and ground for every other pin.

External cabling often acts as antennas for common mode and differential mode noise. This is due to cables often being very long, and numerous. Often times these cables carry communication signals which can be very fast switching digital signals consisting of multiple harmonics. External cabling can be a problem for both conducted and radiated EMI, there are solutions however to reduce problems.

To attenuate common mode noise in external cabling, a ferrite should be placed around the cable. Special ferrites are made for these purposes which have toroidal geometry. This toroidal geometry allows for the ferrite to be clipped on during last minute testing. Two ferrites should be used on each cable and placed near the end of the cable. This attenuates any conducted interference from the device so it is not radiated by the cable, and attenuates any interference induced on the cable from ambient EMI. Figure 2.26 shows an example of a ferrite placed around a cable.

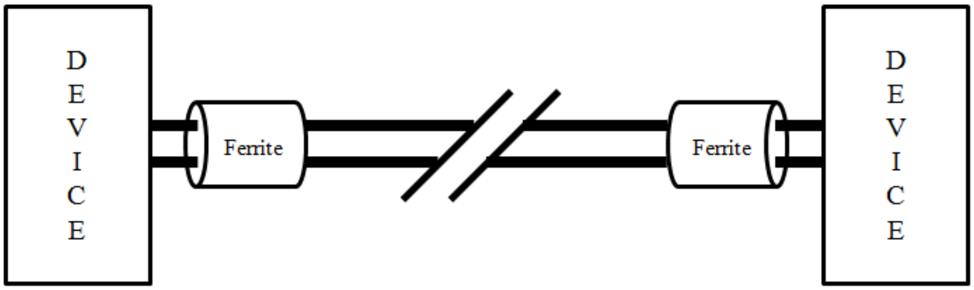


Figure 2.26. Ferrite used for external cabling.

Placing ferrites on each end of the cable also prevents the ferrite from being placed in a current “null”. To attenuate differential mode in external cabling, the wires should be twisted together. Twisting the wires minimizes the effective loop area. This reduces the chance for emissions along with reducing the cable to EMI susceptibility. After implementing these solutions if a radiated emissions problem is still present a shielded cable may be necessary. Shielded cables come in many forms. Some cheaper shielded cables may have a small amount of foil wrapped around the conductors. Other shielded cables may have foil and shielded wire braids wrapped around the conductors. Some shields are more effective than others; however, all act as a barrier to EMI.

After implementing the previous design considerations some radiated EMI issues may still exist. The next approach is to use shielding techniques in the enclosure. The most common form of shielding is to place a metal box around the device. Books have been published on shielding theory alone, and far surpass the expectations of this work. Tips on proper shielding will be presented without going too far in depth.

When shielding enclosures, all seams and gaps must be sealed properly. A metal enclosure can be placed around the radiating device. Ideally, there should be no opening larger than 1/10 of a wavelength of the fastest operating frequency. Once the enclosure is fabricated conductive adhesive tapes can be used for localized shielding around openings or seams. The type of material and thickness also play important roles when shielding. A metal enclosure that is properly grounded will terminate any electric field, but not the magnetic field. Materials with high permeability will provide better shielding for magnetic fields by providing a more conductive return path than air. Depending on the intensity of the fields a thicker enclosure may be needed. The faster the EM wave frequency and higher the field intensities, the farther an EM wave is able to penetrate the skin depth of a shield.

With a properly designed circuit board, external compensation used on cabling, and shielding techniques implemented the final EMI solution is to use a power supply filter. Interference can be picked up on the power grid due to ambient weather conditions or other devices connected to the power grid. The device must be protected from this EMI, and also from EMI generated by the device traveling down the power cord. The power cord is also capable of acting like an antenna. Fortunately, this solution is also the easiest to implement.

2.5.6. Filters

Power supply filters are used to attenuate conducted EMI from the power grid. The power supply filter also prevents interference from the device using the power cord as an antenna. Most commercially purchased AC/DC power supplies all have built in EMI filters. However, if the product has its own power supply design it may not. Power supply filters can be purchased, and are typically housed in the external power cable assembly. Implementing a power supply filter is a very fast and easy way to reduce EMI emission problems.

When designing for EMI, it is always best to assume the worst case scenario. By designing for EMI and taking the necessary precautions it can save thousands of dollars in testing and redesign of the product. It is always easier to initially pass EMC requirements, and then remove unnecessary components to cut cost, rather than build up the product to get it within compliance.

2.6. EMI Simulation Software

In order to reduce time-to-market and remove some of the headaches involved with repeated EMI measurements, it would be nice to have simulation software in the design phase. More specifically, the simulation software should be implemented into the layout software. As the layout complexity increases due to the number of nets in a design manual checking may become extremely time consuming and impractical. An auto-checking tool could alleviate some of these issues. However, as the board complexity increases, the time spent checking the layout for issues will also increase. Once the tool finishes checking it is likely that several nets/routes will have to be corrected and rerouted. The process then repeats and it is likely that software will find issues related to the new layout. This can result in a never-ending loop and due to the complexity of EMI the auto-checking tool itself is unlikely to be able to catch all issues. There are trade-offs between time spent using the EMI simulation software and time spent in the lab performing measurements.

Currently, there are several software tools which are able to be interfaced with the common layout software such as Orcad, Visula, Mentor Graphics. These software tools consist of “Design Adviser”, “Presto”, “Quantic EMC”, “SimLab” and others. These software packages use different philosophies to check for EMI issues. For example, Design Adviser uses rule-checking to verify trace separation, guard-traces, impedance matching, VCC decoupling and isolated copper areas. Presto and Quantic EMC use signal analysis to predict radiated fields. Other more precise software packages exist such as Ansoft HFSS which derive radiated fields from geometry structures using Maxwells equations. Software packages such as HFSS are more suited for analyzing EMI from single components, heat sinks, or shielding analysis. Although HFSS would provide the most accurate radiated field simulation, it simply becomes impractical to model an entire PCB in HFSS.

Due to the overall complexity of EMI, simulation software is not yet where it needs to be to accurately predict radiated emissions. As discussed previously, software is commercially available to simulate radiated emissions from the PCB layout, however this software isn’t able to predict effects due to the enclosure, cabling or other environmental conditions the PCB will be subject to. Very precise EM modeling tools such as HFSS are not able to model complete circuit boards yet, and the computing power needed to do so would take a very long time.

2.7. Flowchart for Troubleshooting EMI

A flowchart is presented in order to aid in troubleshooting devices during the EMC testing process. This flowchart assumes that the DUT has failed the radiated emissions test, and begins by isolating problem areas to cabling or the equipment box. The flowchart then attempts to further narrow down problem areas and suggest improvements which could be made. Figure 2.27 shows a process flowchart for troubleshooting EMI.

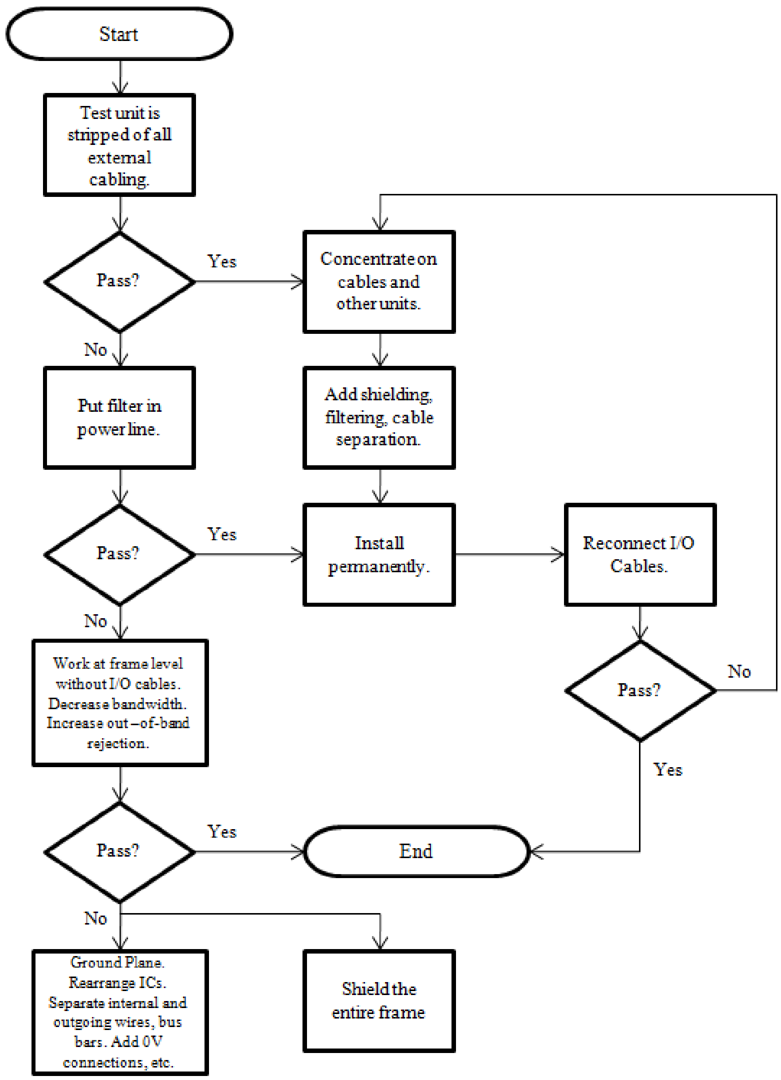


Figure 2.27. EMI Troubleshooting Flowchart (Mardiguian, 2001).

The flowchart (Figure 2.27) first suggests that all external cables be removed. Once the cables are removed, the test should be run again to see if the EUT passes. If the answer is yes a focus can be made on reducing radiated emissions in the cabling. Methods presented in the solutions section such as shielding, filtering, or seperating the cables can be done. The cables can then be reattached and EUT can be tested again. If the EUT fails even with the cables disconnected two assumptions can be made. The power cord is radiating, or the equipment box is radiating. A power line filter should be installed to filter out common-mode and differential-mode noise on power line, and the EUT tested again. If the unit fails again the equipment box is left as the primary source of emissions.

At this point it is understood that the EUT has an emissions source that is originating from the equipment box. Prior to making any changes a sniffer probe should be used to scan all faces of the enclosure making sure to capture the output for further analysis. The measurements taken by the probes have no significant meaning, but can be used to track improvements. A reduction in the measured near field emissions of the probe (ΔdB) should provide roughly the same amount in the far field during testing. Using the probes, particularly noisy areas around the equipment box can be identified. Specific measures can then be implemented such as shielding the entire frame, adding ground planes or additional ground connections, proper PCB zoning, minimizing loop areas, minimizing critical trace lengths, and reducing the other interference coupling paths mentioned in this chapter.

This chapter discussed the basics of EMI beginning with electromagnetic theory and propagating waves. EMI definitions, measurement techniques, standards, problems, solutions and a flowchart to assist in troubleshooting methods were all presented. The next chapter will discuss an experiment on EMI by designing circuits which intentionally exacerbate EMI, and then implement many of the reduction techniques discussed in this chapter.

CHAPTER 3. EMI MEASUREMENTS

In attempting to break down complicated EMI problems with different electronic devices, multiple circuit designs were analyzed separately. By analyzing each problem independently observations could be made about how each contributes to the overall EMI problem. Since EMI is such a dynamic and situational problem, broad and general observations were made in order to make predictions that will hold true for multiple scenarios. Finally, a case study is evaluated involving a mixed analog and digital circuit board which would better represent a consumer device that would have to pass FCC EMI requirements.

All radiated EMI measurements in this chapter were made using a $\vec{H}$ field probe designed by Com-Power Corporation. The $\vec{H}$ field probe has a frequency range of 0.3MHz - 100MHz. A wooden jig was made which fixed the height of the probe 25.4mm above the equipment under test (EUT). The probe was connected to a HP 8447D pre-amp which has a fixed gain of 25dB between 0.1MHz - 1300MHz. Measurement data was captured with an Agilent spectrum analyzer model N1966A. Unless otherwise specified, all measurement data was taken between DC-1GHz which is the FCC specification for radiated EMI measurements.

3.1. Switching Regulator

Power management is a critical aspect of any design. Efficient power conversion is especially important in digital systems where the input voltage is significantly higher than the logic voltage. Switching regulators are used in many designs over linear regulators primarily due to their efficiency. However, due to the basic operation of the switching regulator it generates an excessive amount of noise over the basic linear regulator. Switching speeds varying from 50kHz-250kHz and the switching waveform is close to a square wave. Due to these characteristics, the switching regulator produces an excessive amount of noise and harmonics.

The unwanted noise and harmonics produced by a switching regulator can be minimized using filters and using good layout techniques. Implementing ferrites and minimizing trace lengths are critical to reduce EMI. The amount of radiated EMI is also dependent on whether common-mode or differential-mode interference is present. A circuit board was designed consisting of four switching regulators. Two regulators were designed to reproduce common-mode interference, and two were designed to reproduce differential-mode interference. Figure 3.1 shows the differences in the regulator schematics between common-mode and differential mode interference.

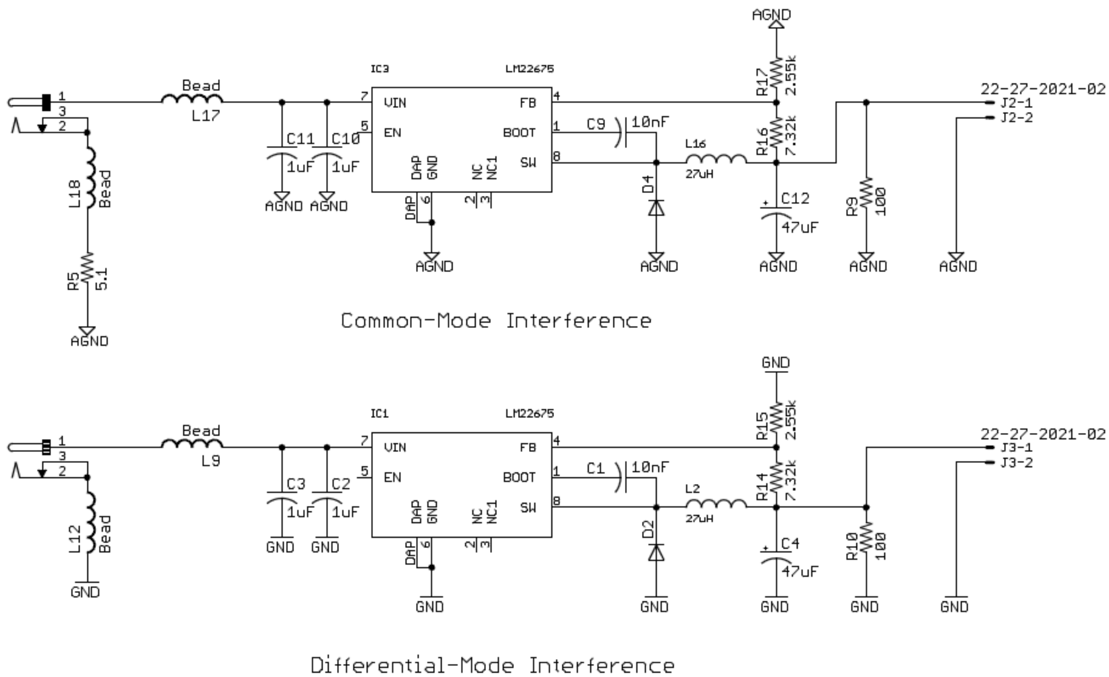


Figure 3.1. Switching Regulator Schematic.

The only difference in the schematic between the common-mode and differential-mode examples is the resistor (R5) located next to the DC barrel connector. The purpose of this resistor is to create multiple current loops. For the common-mode example there is a ground loop from the load resistor to the switching regulator, and also from the switching regulator to the source. This resistor is used to simulate improper grounding techniques where multiple return paths may exist that have different impedances.

Of the designs to represent common-mode and differential-mode interference, one regulator from each group followed good layout techniques, and the other had a poor layout. The circuit board also had a load which could be attached in close proximity to each switching regulator, or five unique loads with different types of return paths to measure the effects of the conducted EMI. Figure 3.2 shows the top copper and silkscreen of the switching regulator PCB.

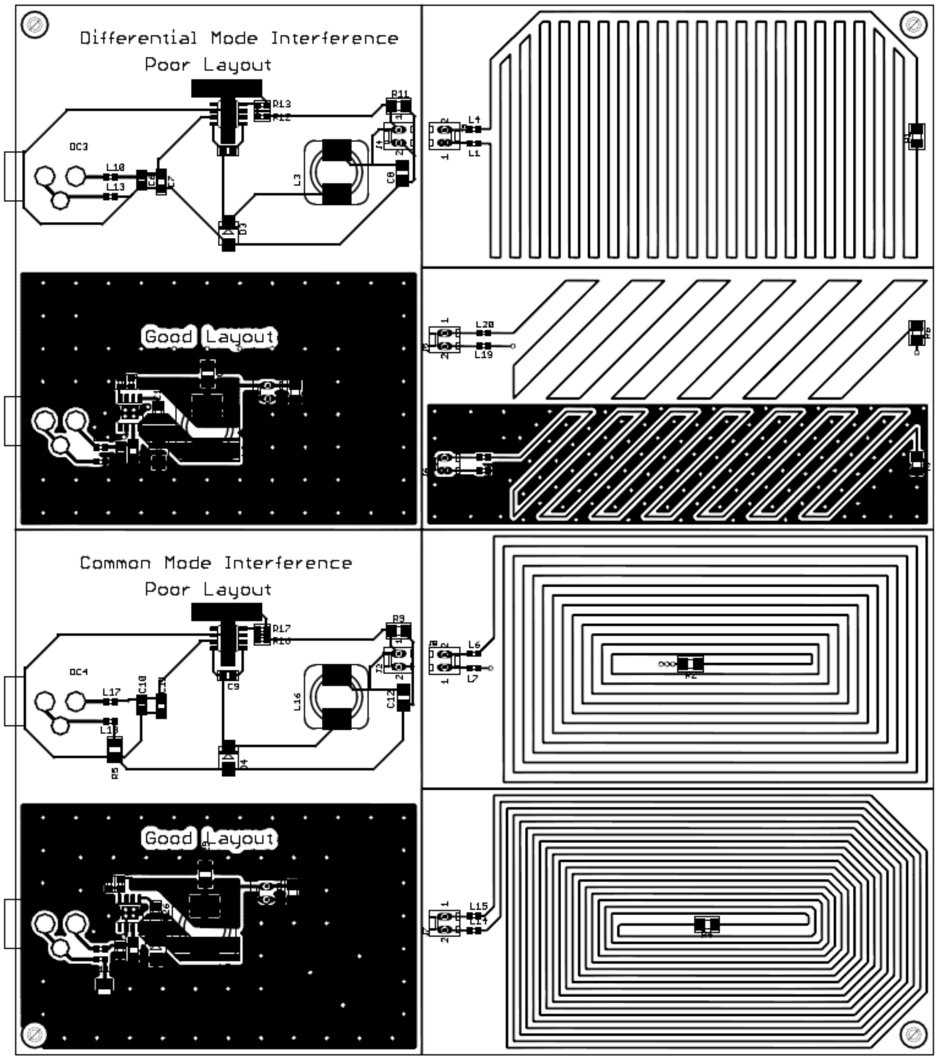


Figure 3.2. Switching Regulator PCB.

From the switching regulator PCB layout (Figure 3.2) the good layout has large copper planes shared by many components, and all of the circuitry is kept very compact and in close proximity to one another. There is also a large ground plane surrounding the components, and directly underneath them. The poor layout places the components far away from one another, effectively maximizing loop areas. Traces are also very narrow and much more inductive than the large planes used in the good layout. The poor layout also has no ground planes which forces current through the high inductive traces and loop areas. The five unique return paths, from top to bottom, are numbered one through five respectively. Table 3.1 describes each unique load.

Table 3.1.
Attached Load Descriptions.

Load Number	*Description*
1	Short source trace length, long complex return path
2	Return path is directly under source
3	Identical to Load2, except with a copper pour for the return path
4	Long complex source trace length, short return path
5	Complex source and return path running parallel to one another

The load resistance was kept constant for each regulator at 100Ω, providing 50mA of current through each load route. Measurements of the radiated EMI were taken of each switching regulator with the load in close proximity, and with each of the five unique loads attached. When taking measurements the probe was placed directly over the load resistor. Different ferrites were then installed in the circuit at the input and output of the switching regulator and the measurements repeated.

3.1.1. Differential Mode Interference

Measurements were first made with the load placed close to the regulators designed to simulate differential mode interference. Minimizing trace lengths between the load and the switching regulator insure that all measured radiated emissions are from the switching regulator itself, and not due to antenna effects from conducted EMI. The effects of ferrites at the input of each regulator were measured first. Measurements were made first with no ferrites, then with 100Ω and 600Ω ferrites. Figures 3.3, 3.4 and 3.5 show the measurement data for each ferrite configuration.

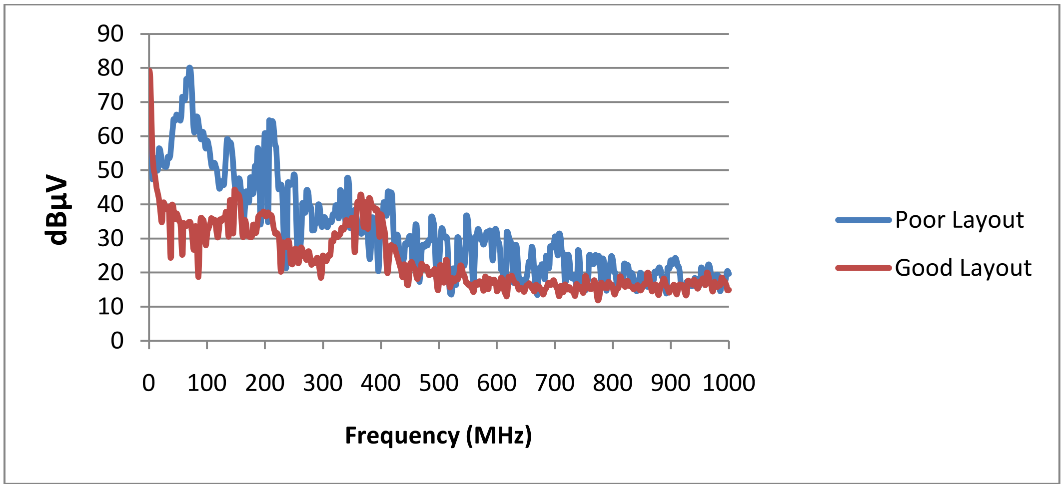


Figure 3.3. Differential Mode No Ferrites.

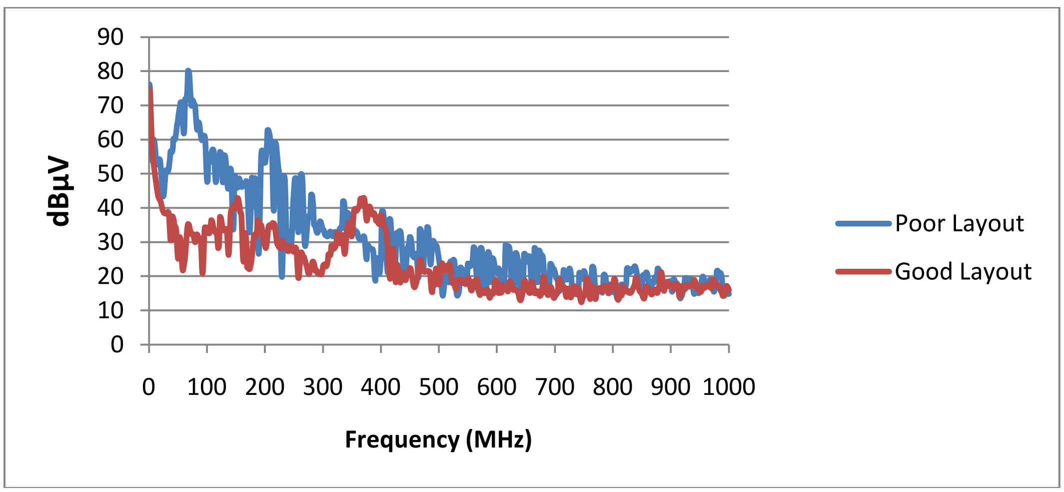


Figure 3.4. Differential Mode 100Ω Ferrites.

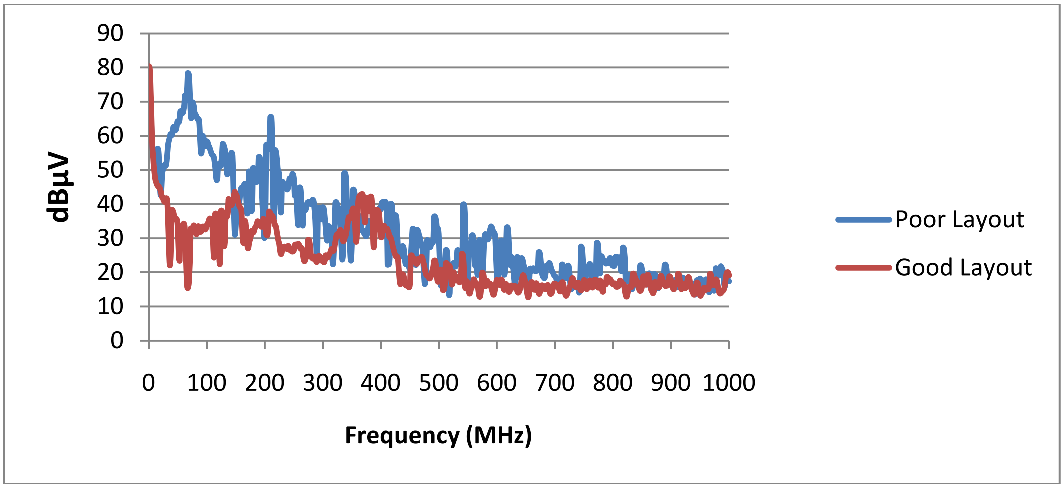


Figure 3.5. Differential Mode 600Ω Ferrites.

From the data it is obvious that the effects from the ferrites at the input are negligible. This is due to the DC power supplied to the regulators being clean. No ferrites were needed on the output of the regulator since the load was placed in close proximity. The importance of this data shows that a good layout will significantly reduce the amount of radiated emissions from a switching regulator. Observing the data at 70MHz it can be seen that there is a difference of close to 60dB in radiated emissions between the good and poor layout. Next, five loads are next attached with return paths designed to simulate both correct and improper routing on a PCB.

3.1.2. Differential Mode Attached Loads

Load1 is attached to the differential mode switching regulators and examined first. Load1 had a short source trace length, and long complicated return path. From the data (Figures 3.6 through 3.8) it can be seen that the 600Ω ferrites were more effective in reducing emissions than the 100Ω ferrites or no ferrites configuration. A drop of approximately 5-8dB was achieved by using the 600Ω ferrites.

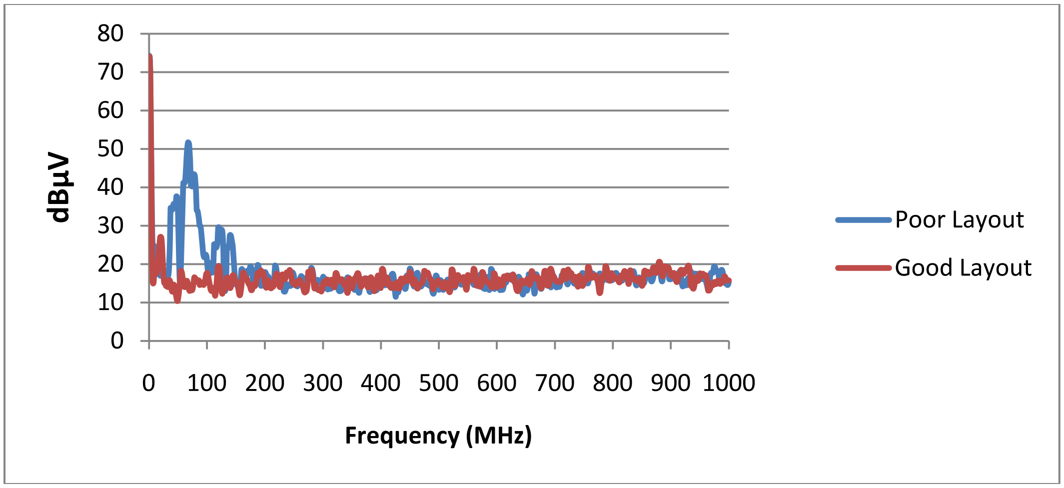


Figure 3.6. Differential Mode Load1 600Ω Ferrites.

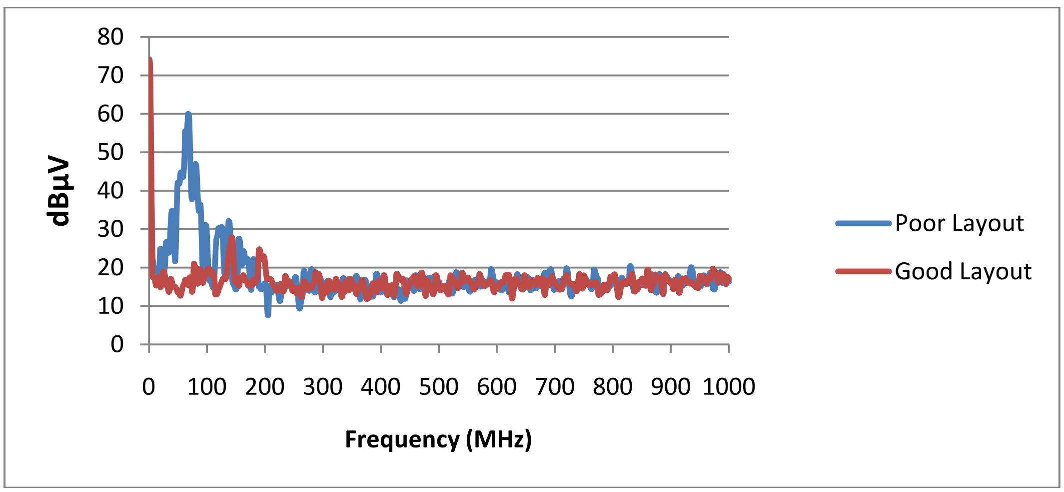


Figure 3.7. Differential Mode Load1 100Ω Ferrites.

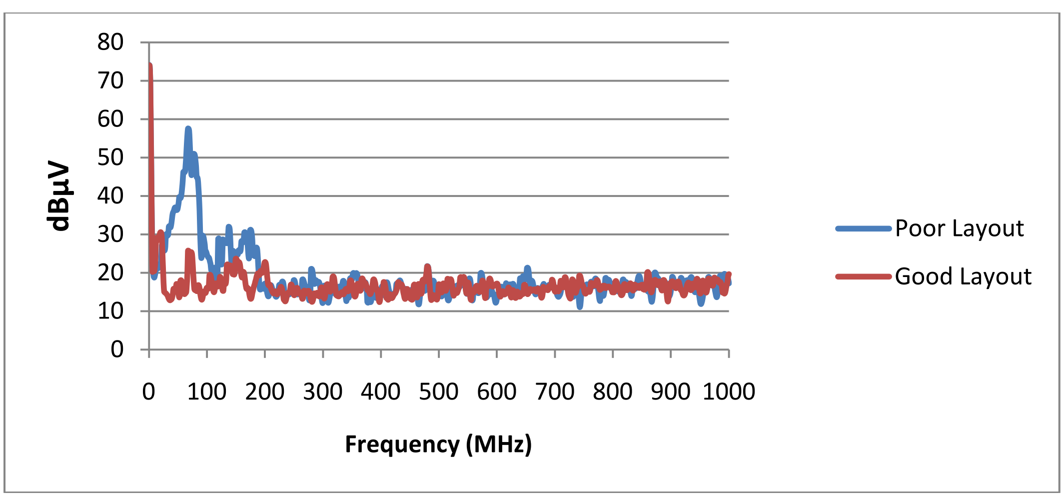


Figure 3.8. Differential Mode Load1 No Ferrites.

Load2 was designed to be proper routing with the return path directly under the source. When attached to the differential mode regulator with a good layout load2 has the least amount of radiated emissions overall, even with no ferrites installed. When the differential mode regulator with the poor layout is used, load2 radiates approximately 30dB more than the good layout at approximately 70MHz in all measurements. Figures 3.9 through 3.11 show the differential mode measurements for load2.

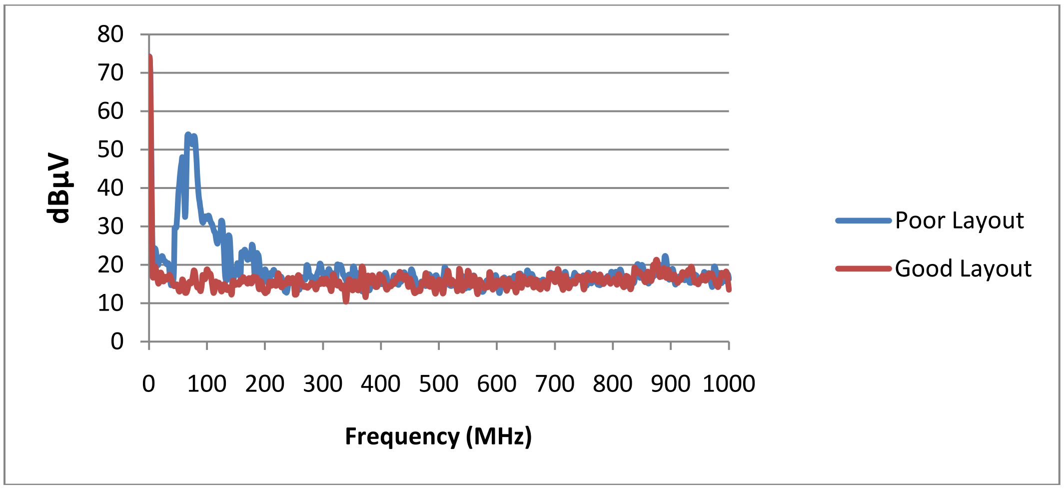


Figure 3.9. Differential Mode Load2 600Ω Ferrites.

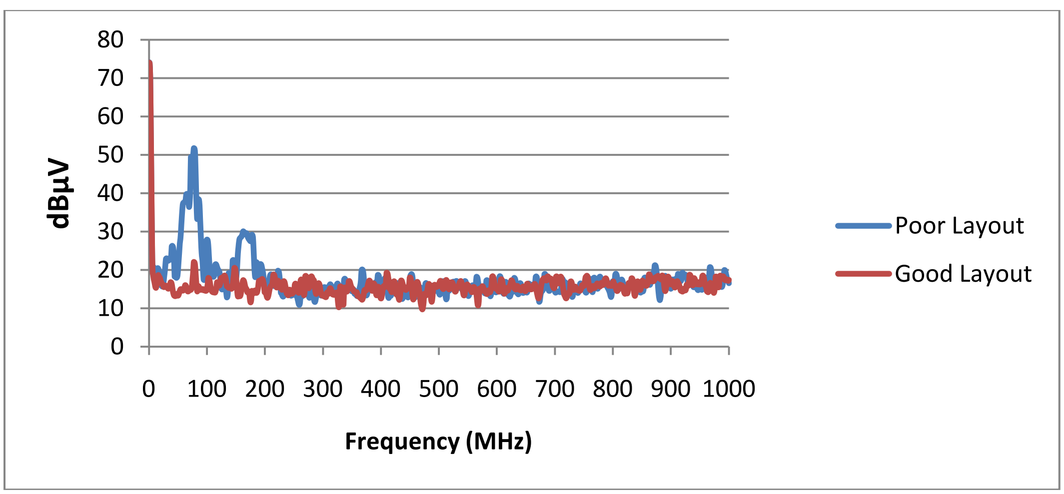


Figure 3.10. Differential Mode Load2 100Ω Ferrites.

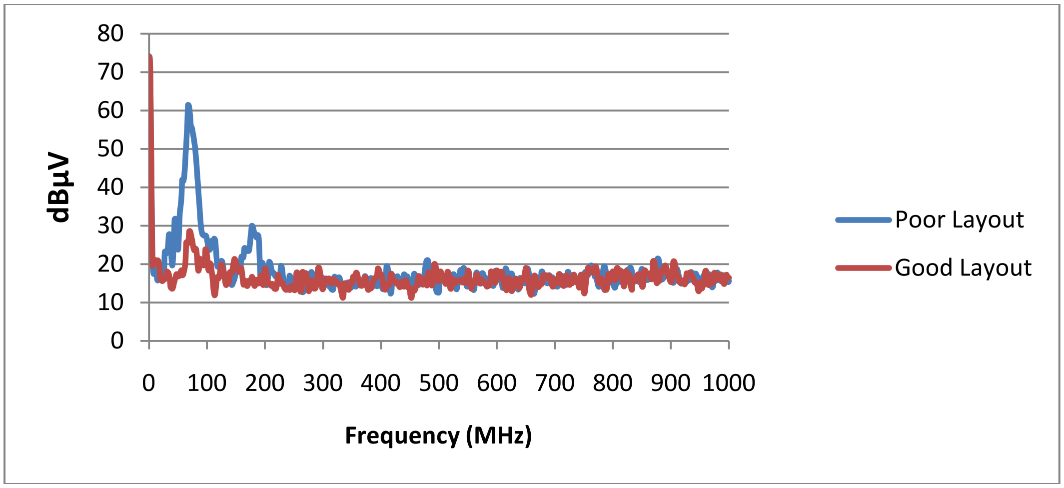


Figure 3.11. Differential Mode Load2 No Ferrites.

Load3 is identical to load2, except that a large copper pour was used for the return path instead of a single trace. This is also considered good routing as the return path is unobstructed, and can follow a path directly beneath the source current. When comparing load2 and load3 measurement results the 600Ω, and 100Ω configurations are almost identical. However, load3 with no ferrites radiated roughly 10dB more than load2 with no ferrites. Figures 3.12 through 3.14 show the measurement data for load3.

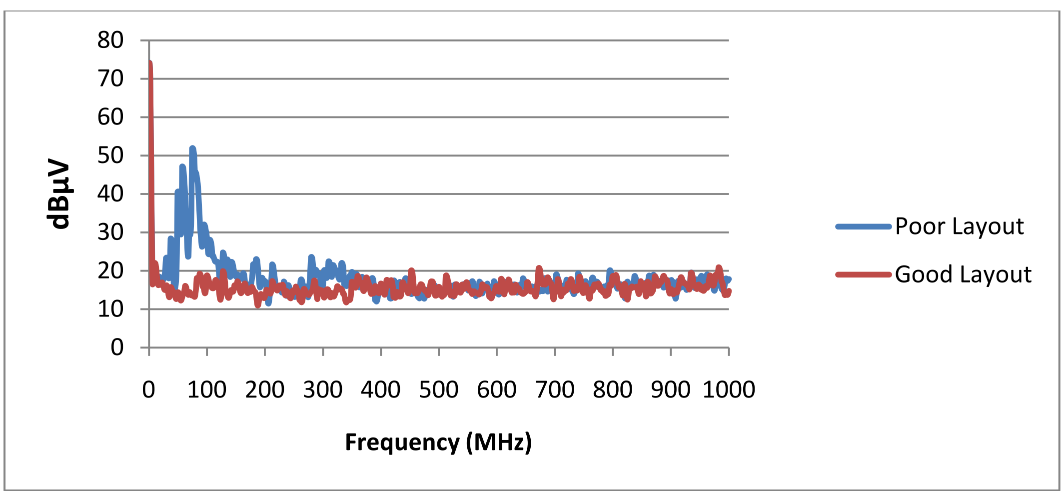


Figure 3.12. Differential Mode Load3 600Ω Ferrites.

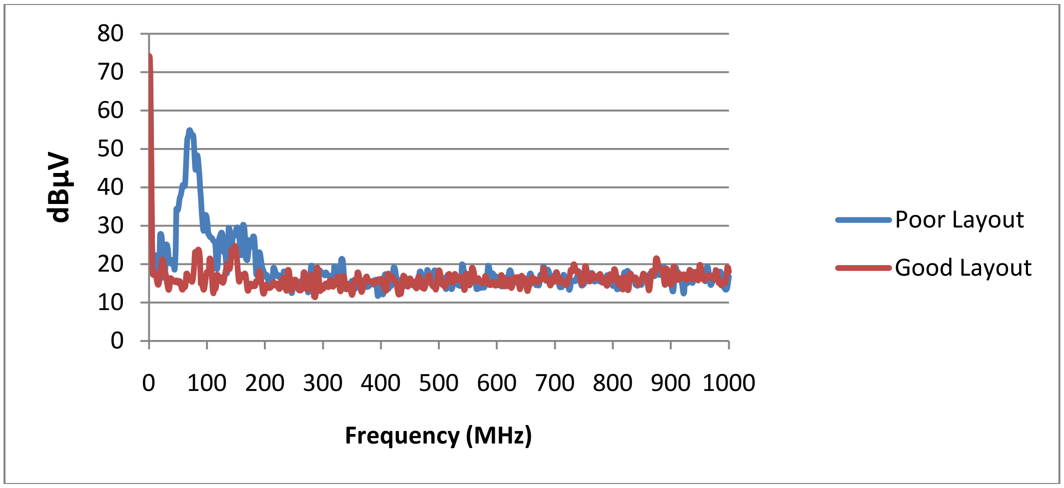


Figure 3.13. Differential Mode Load3 100Ω Ferrites.

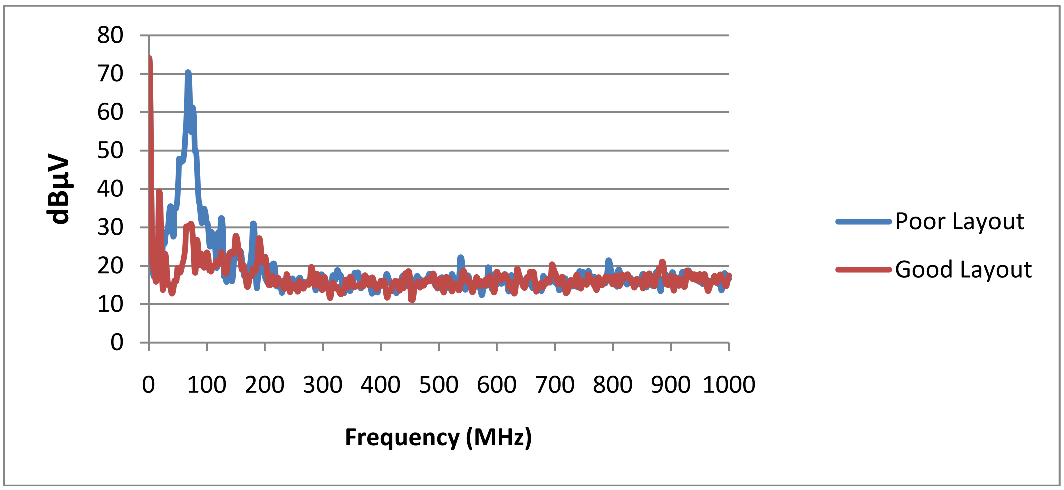


Figure 3.14. Differential Mode Load3 No Ferrites.

Load4 is the opposite of load1 with a long complex source trace and short return path. The 600Ω ferrite configuration with load4 yielded the lowest radiated emissions. Observing figures 3.15 through 3.17, the poor regulator layout at 70MHz with 600Ω ferrites reduced emissions by 7dB over the 100Ω configuration, and approximately 17dB with no ferrites.

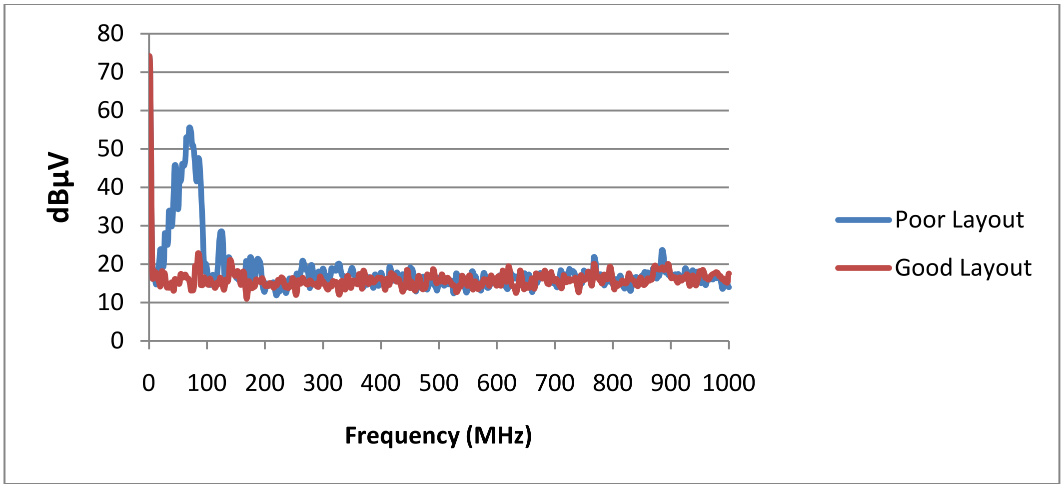


Figure 3.15. Differential Mode Load4 600Ω Ferrites.

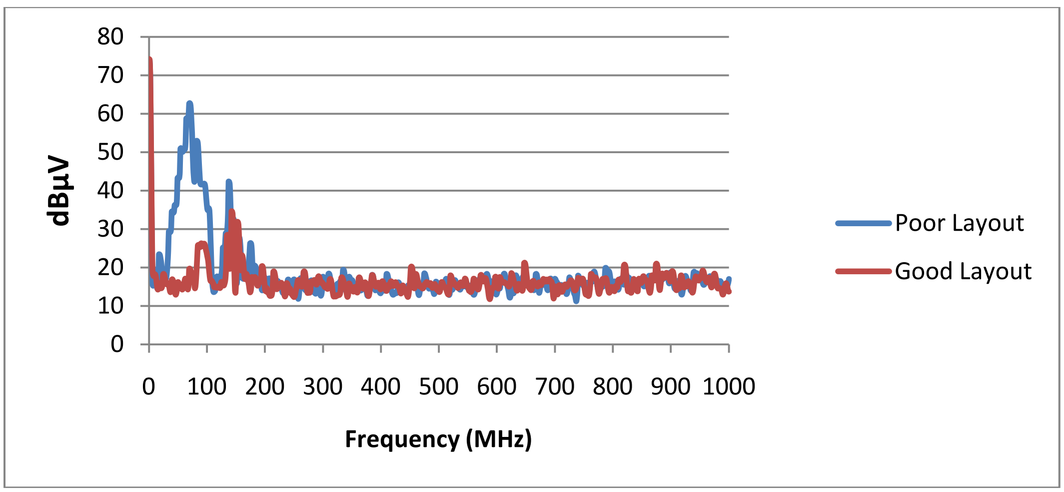


Figure 3.16. Differential Mode Load4 100Ω Ferrites.

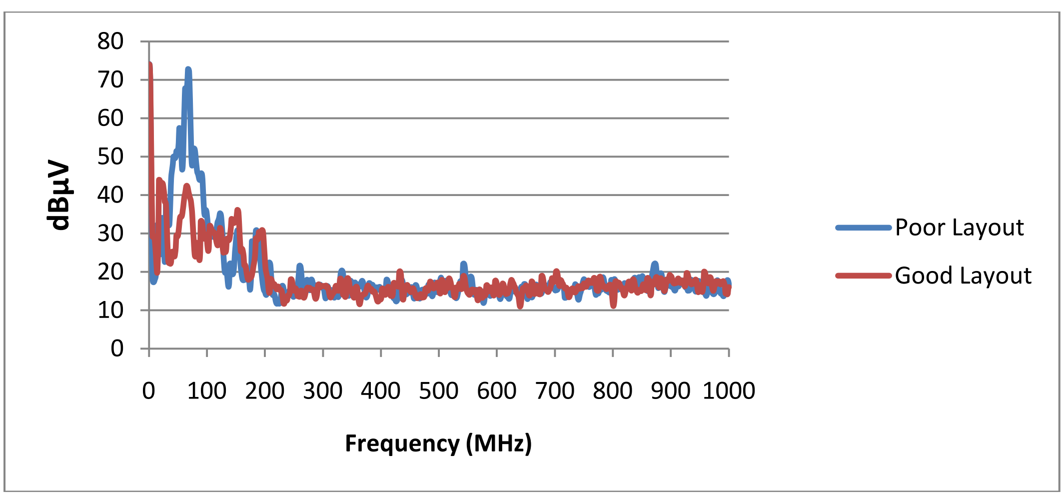


Figure 3.17. Differential Mode Load4 No Ferrites.

Load5 was designed to have the worst routing by having a long complex source and return path. The measurement data for load5 (Figures 3.18 through 3.20) confirms this by radiating a significant amount more than any other load between 50MHz and 250MHz. When paired with the switching regulator with the poor layout radiated emissions approached 80dBμV. Observing the effects of the ferrites on the poor layout, there was no reduction between the 600Ω and no ferrite configuration. When the good regulator layout was used the ferrites were able to further reduce radiated emissions, however emissions were still larger than other loads.

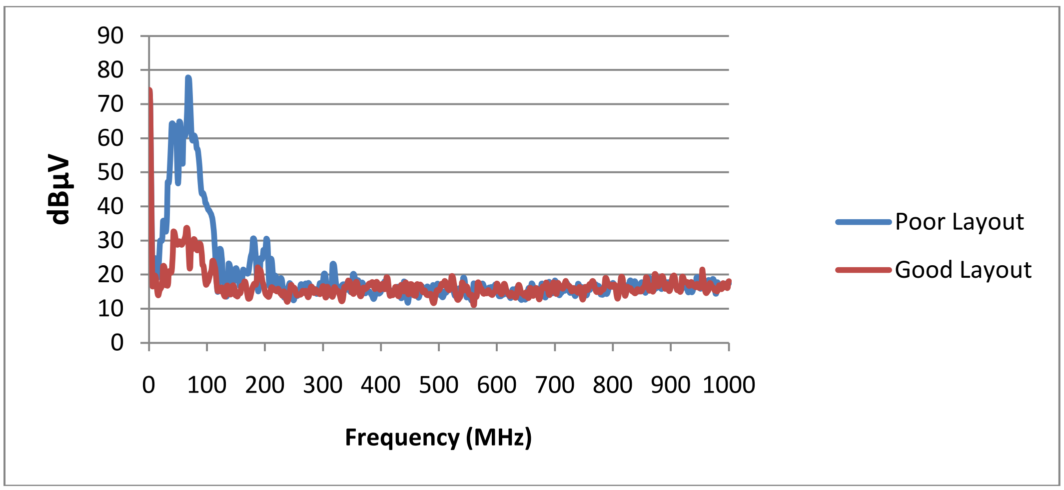


Figure 3.18. Differential Mode Load5 600Ω Ferrites.

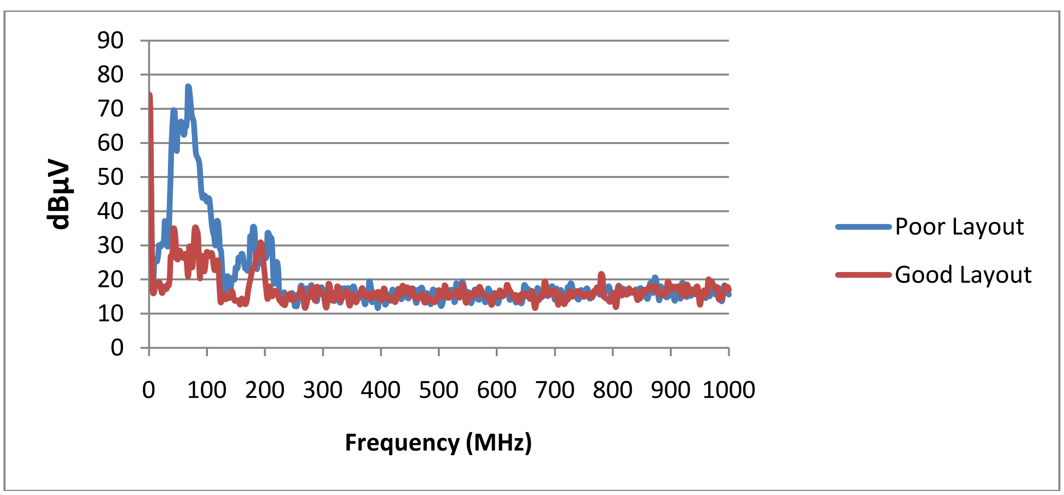


Figure 3.19. Differential Mode Load5 100Ω Ferrites.

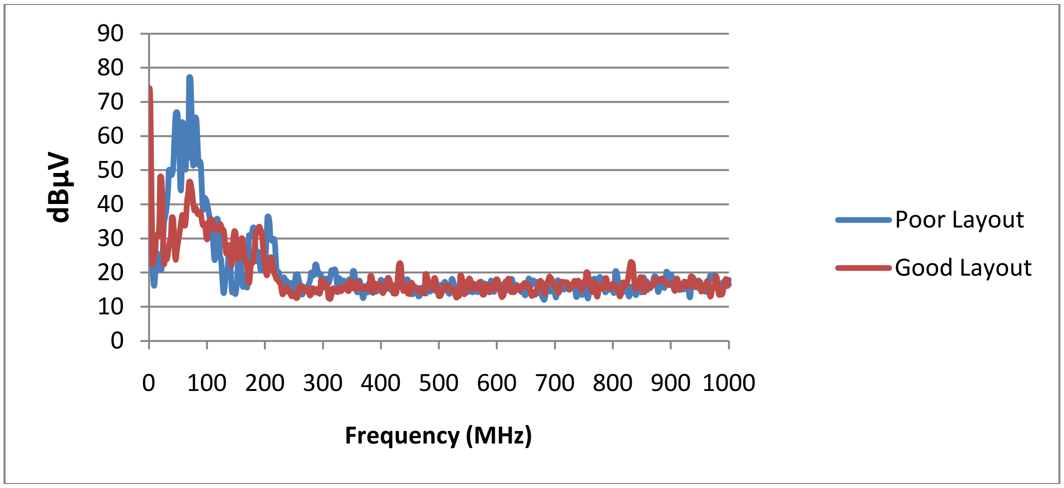


Figure 3.20. Differential Mode Load5 No Ferrites.

Observing the results from Figure 3.6 through 3.20, there is a large difference between each load with different ferrite configurations. Using ferrites, the amount of radiating emissions was reduced, but not completely eliminated. There was a noticeable difference between no ferrites and 100Ω, but only a marginal improvement from 100Ω to 600Ω. It is clear that a good layout will significantly reduce radiated emissions.

3.1.3. Common Mode Interference

As described previously, measurements were first made with the load in close proximity to the switching regulator to rule out any effects of conducted EMI. Since it was observed that ferrites on the input are negligible, only the measurements with no ferrites will be examined. Figure 3.21 shows common mode interference with the load in close proximity to the regulator. Data for the 100Ω and 600Ω ferrites can be found in the appendix.

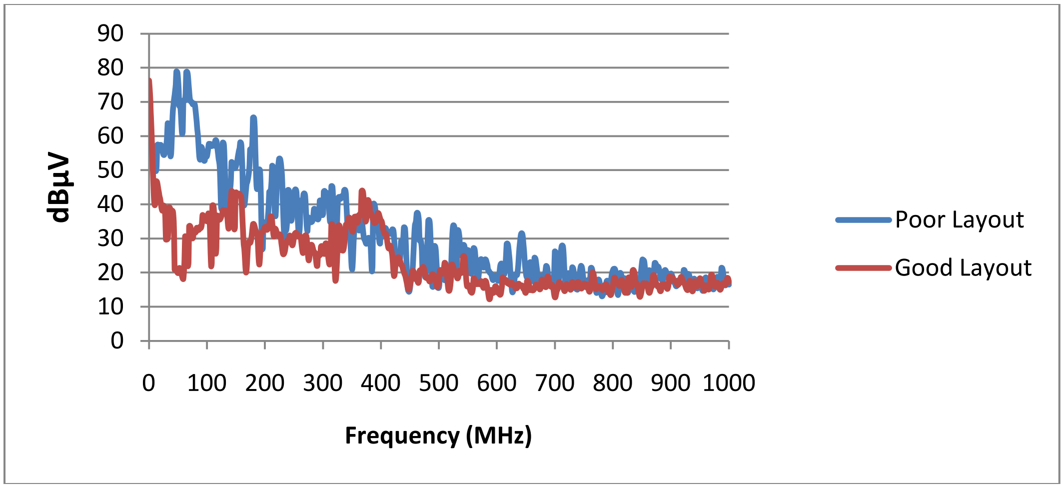


Figure 3.21. Common Mode No Ferrites.

From the data, it can be seen that a good layout is critical to reduce radiated EMI from switching regulators. When comparing the amount of radiated emissions between the common-mode and differential-mode regulators, the regulators using a good layout had no significant change in radiated emissions while the regulators with a poor layout were degraded. This is expected since common-mode interference is due to multiple return paths, but differential-mode interference is unavoidable and still present. A good layout will eliminate these ground loops, while the poor layout will increase the size of the loops. The five loads used previously were attached to the regulators designed to enhance common-mode interference, as will be seen in the next section.

3.1.4. Common Mode Attached Loads

Interference due to common-mode noise is noticeably increased than interference due from differential-mode interference, especially with a poor layout. Load1 common mode measurement results are shown in Figures 3.22 through 3.24. Strangely, the 600Ω ferrites had worse radiated emissions than the 100Ω and no ferrite configurations. When comparing common mode interference to differential mode interference with 600Ω ferrites installed and a poor layout (Figures 3.22 and 3.6), it can be seen at 70MHz that the common mode example has a 30dB increase in radiated emissions.

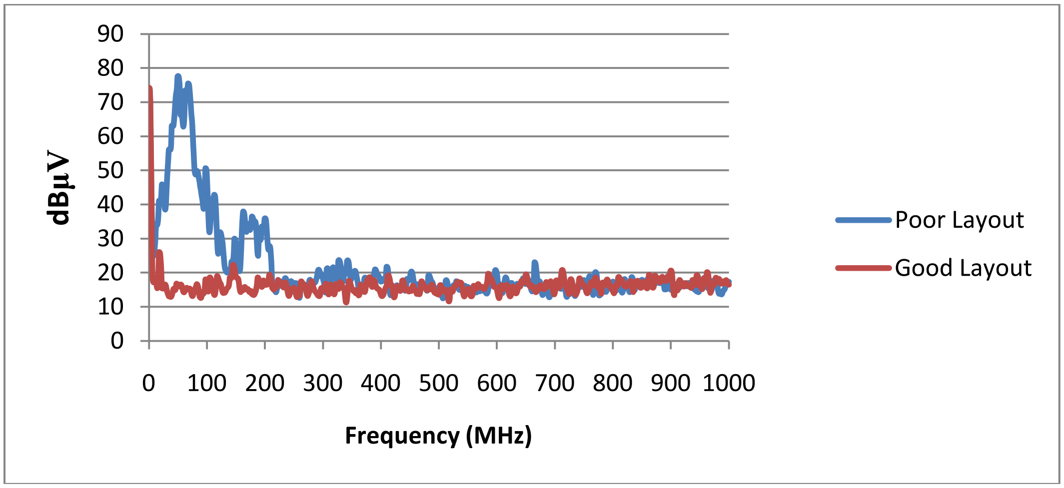


Figure 3.22. Common Mode Load1 600Ω Ferrites.

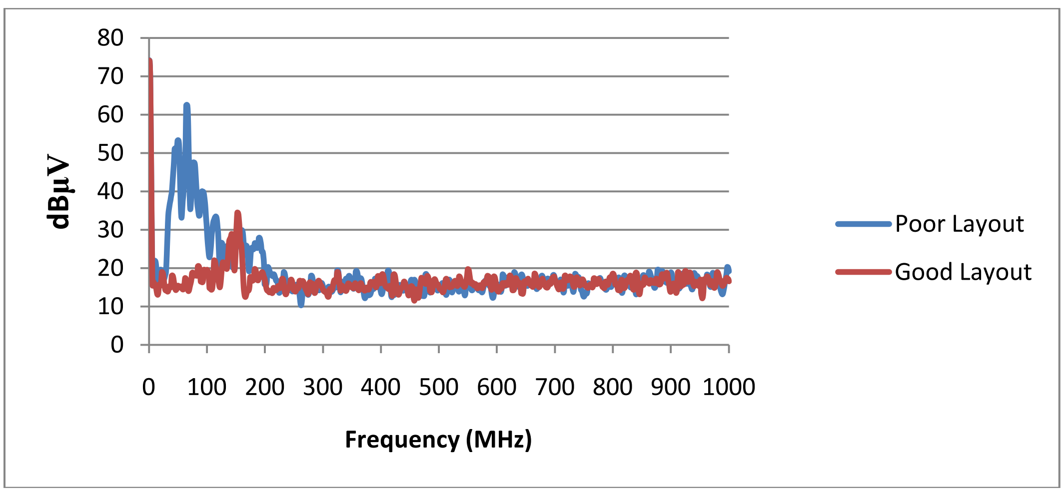


Figure 3.23. Common Mode Load1 100Ω Ferrites.

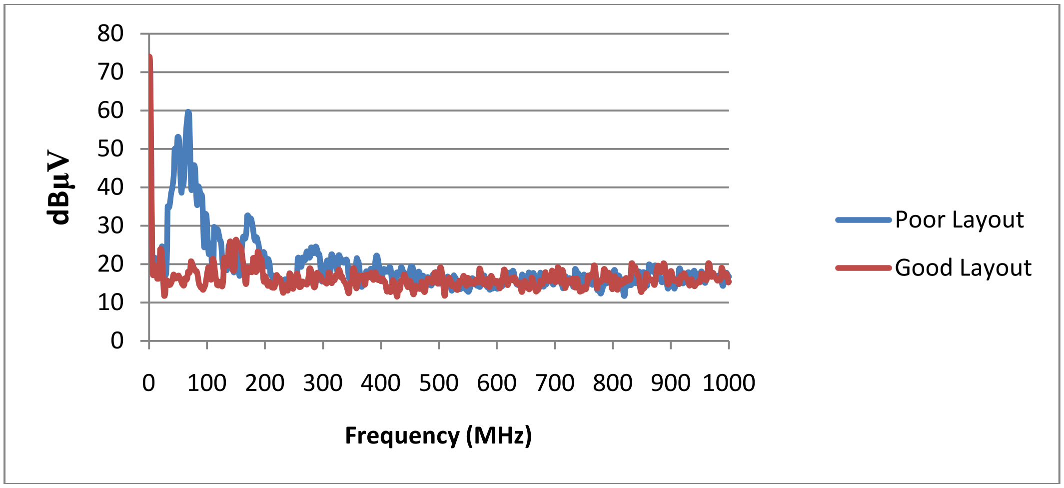


Figure 3.24. Common Mode Load1 No Ferrites.

Load2 measurement results are shown in Figures 3.25 through 3.27. With a good regulator layout load2 showed little to no radiated emissions above the noise floor, even with no ferrites installed. The 600Ω ferrites performed the best by reducing emissions at 70MHz over the 100Ω ferrites by 5dB, and over no ferrites by 35dB.

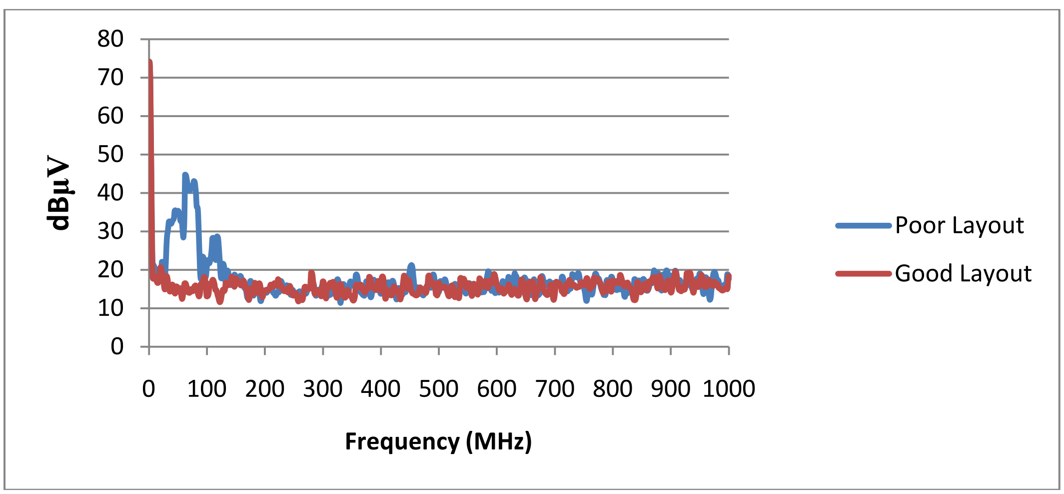


Figure 3.25. Common Mode Load2 600Ω Ferrites.

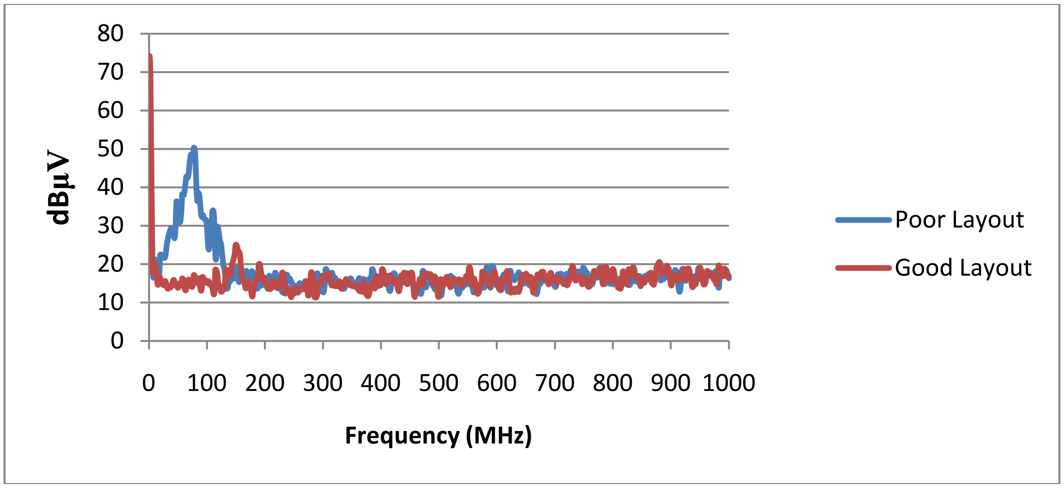


Figure 3.26. Common Mode Load2 100Ω Ferrites.

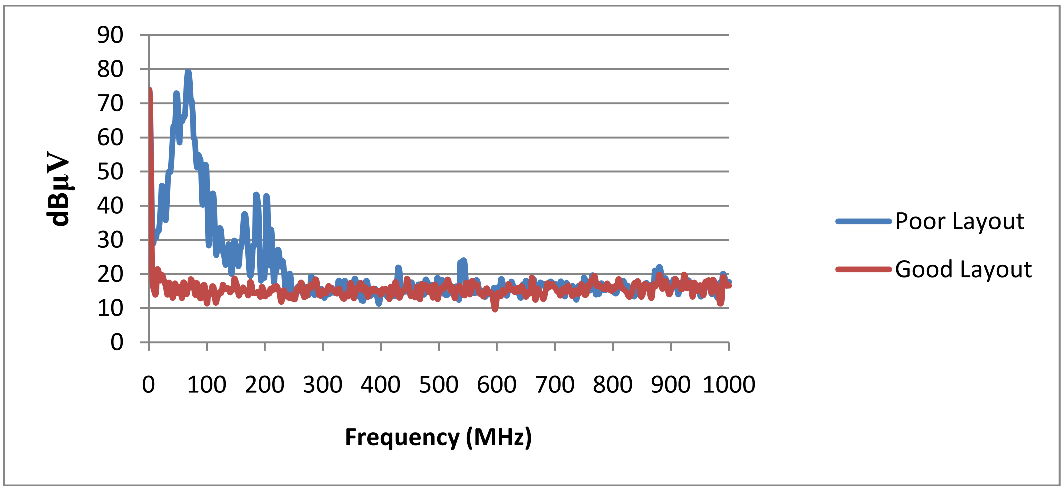


Figure 3.27. Common Mode Load2 No Ferrites.

Load3 had the least amount of emissions with common mode noise. Figures 28 through 30 show Load3 measurement results. There are no radiated emissions with the good regulator layout. With the poor layout there was little difference between the 600Ω and 100Ω ferrites, and only a marginal increase with no ferrites.

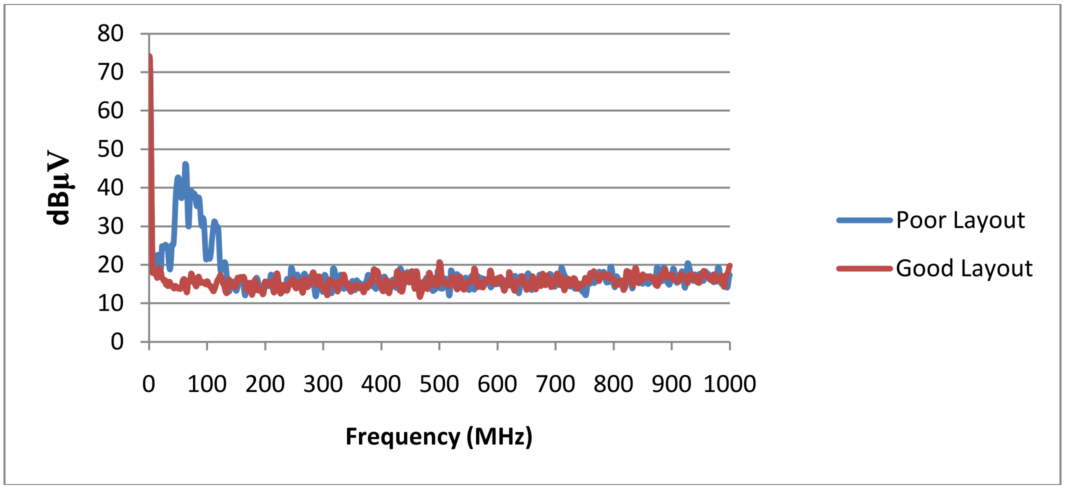


Figure 3.28. Common Mode Load3 600Ω Ferrites.

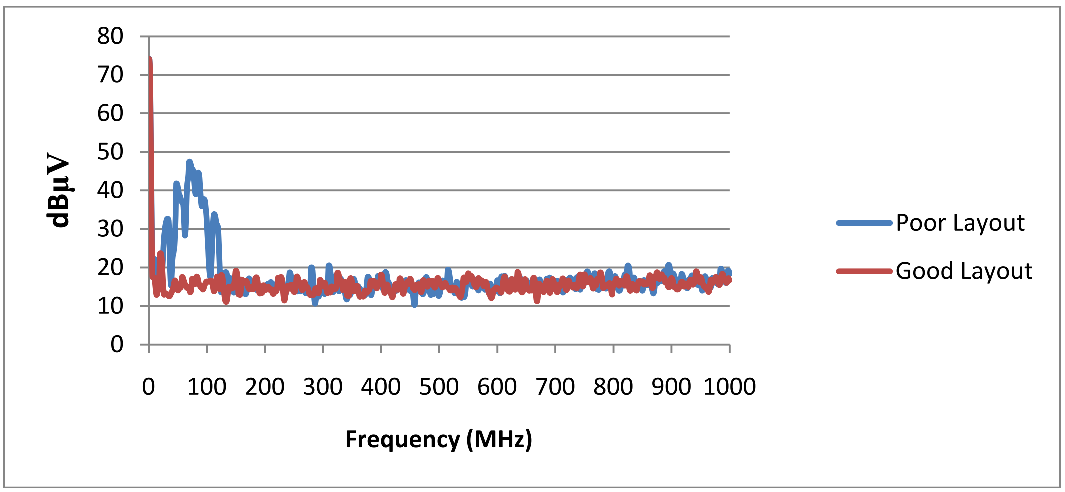


Figure 3.29. Common Mode Load3 100Ω Ferrites.

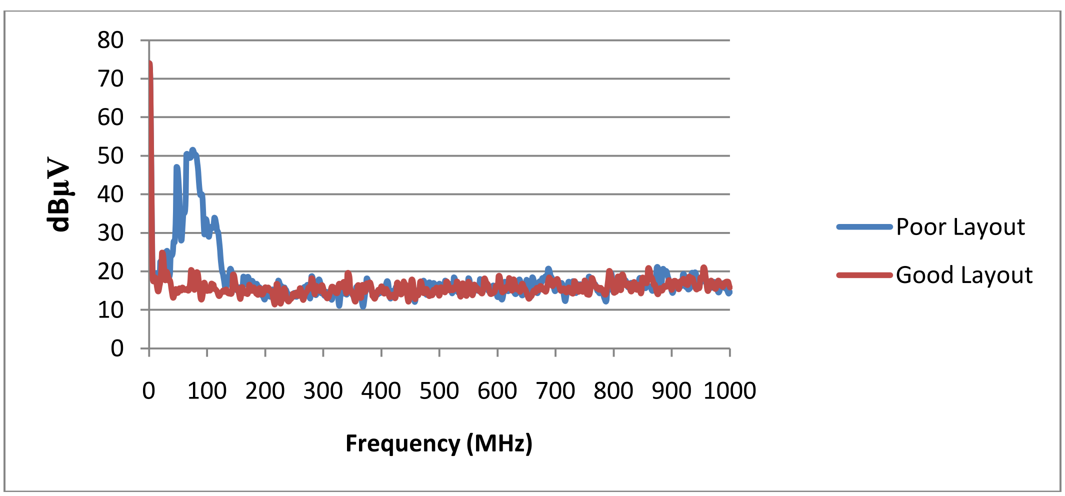


Figure 3.30. Common Mode Load3 No Ferrites.

Load4 measurement results are shown in Figures 3.31 through 3.33. Similar to load1, the poor layout and 600Ω ferrites had higher radiated emissions than the 100Ω and no ferrite configurations at 70MHz. This indicates there may be a performance related issue when using the larger ferrite. The 100Ω ferrites were successful in reducing emissions over the no ferrite configuration.

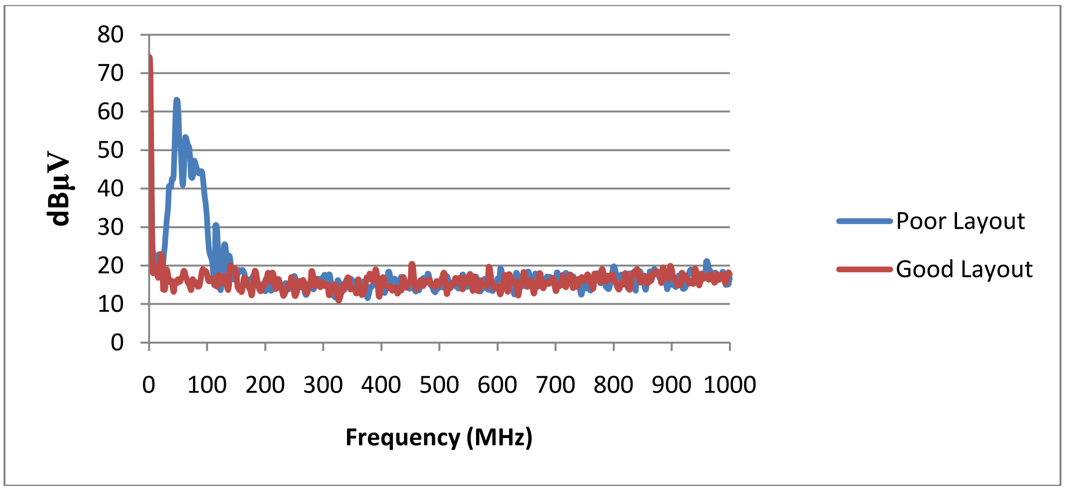


Figure 3.31. Common Mode Load4 600Ω Ferrites.

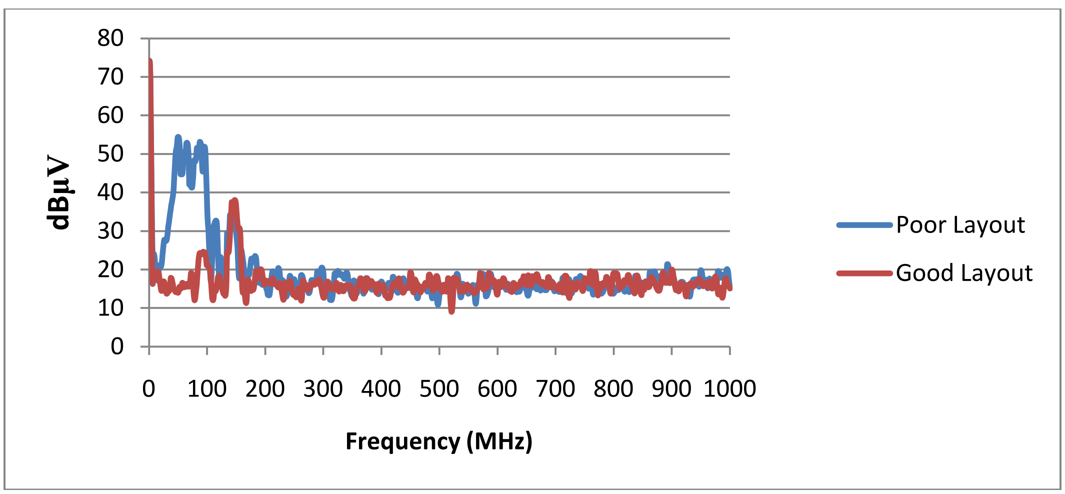


Figure 3.32. Common Mode Load4 100Ω Ferrites.

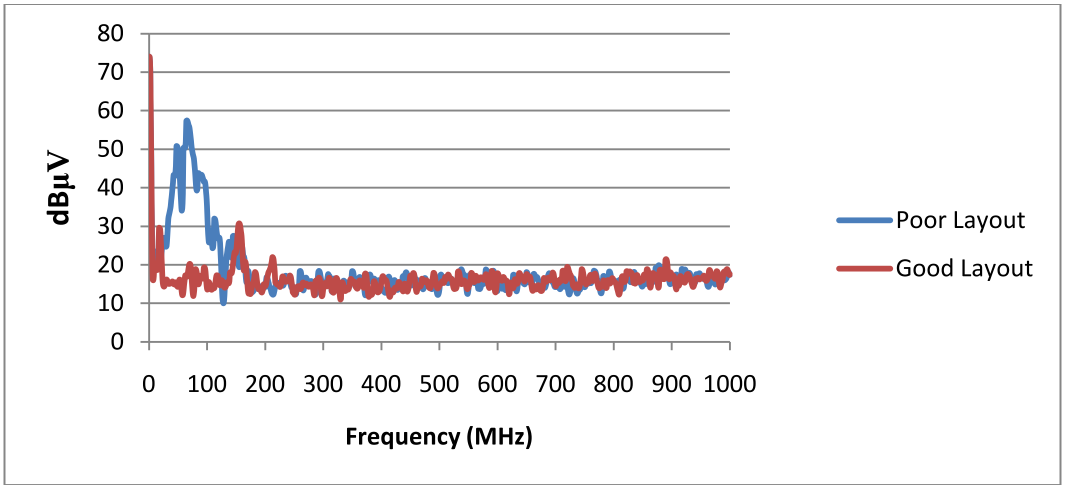


Figure 3.33. Common Mode Load4 No Ferrites.

Similar to the differential mode example (Figures 18 through 20), load5 again had the highest radiated emissions, when compared to the other loads, between 50-250MHz. Load5 measurement results are shown in Figures 34 through 36. Unlike the differential mode example, the 600Ω ferrites were able to reduce the amount of radiated emissions due to common mode noise. The 100Ω ferrites did not have an impact on radiated emissions over the no ferrite configuration.

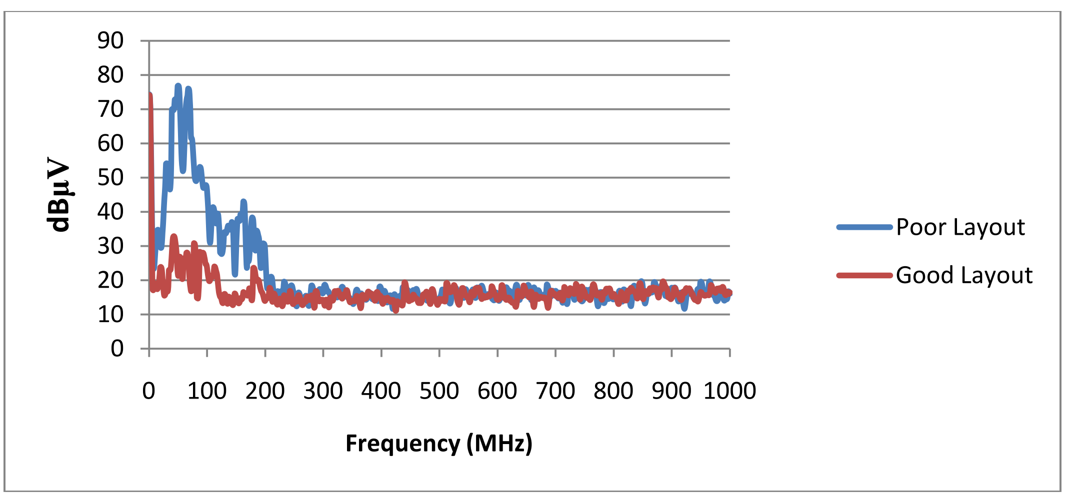


Figure 3.34. Common Mode Load5 600Ω Ferrites.

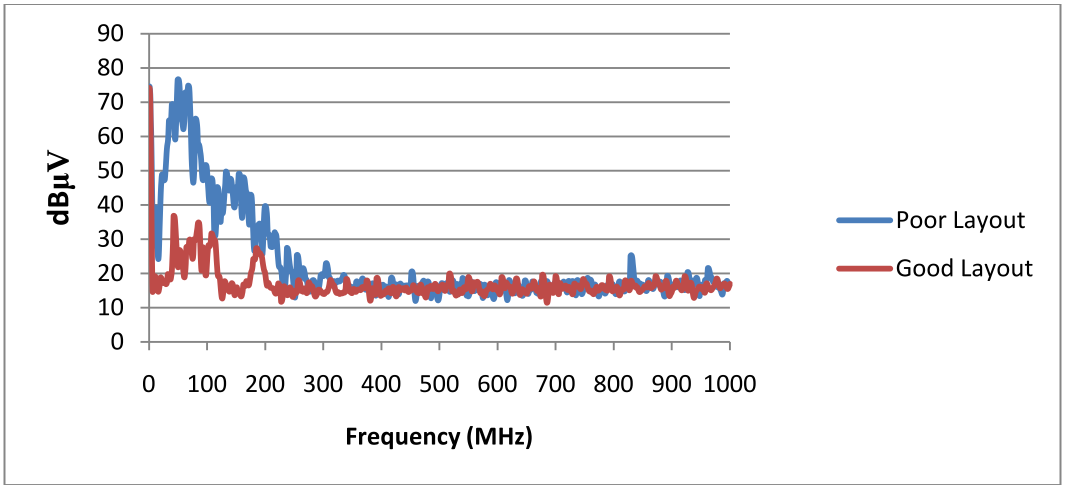


Figure 3.35. Common Mode Load5 100Ω Ferrites.

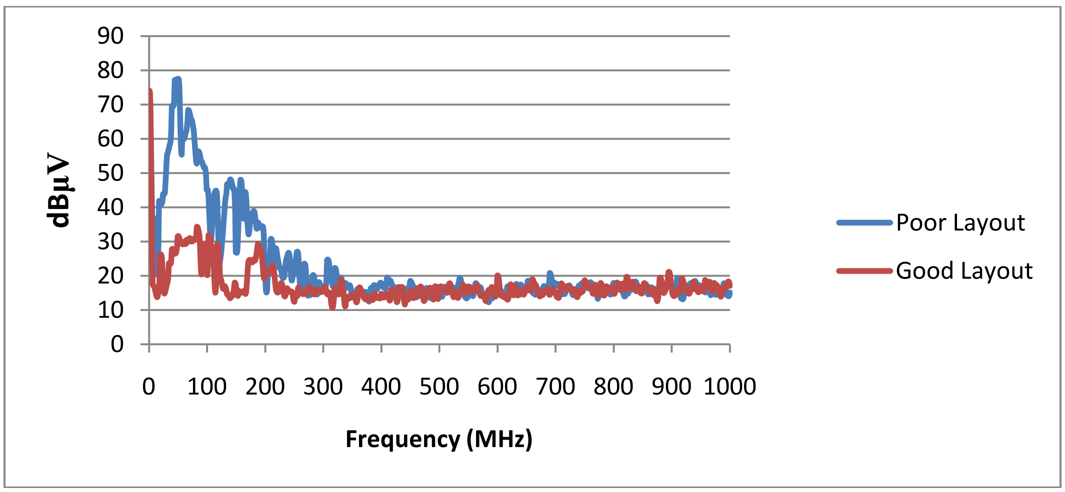


Figure 3.36. Common Mode Load5 No Ferrites.

Evaluating Figures 3.22 through 3.36, it is clear though that noise due to common mode interference can be significantly reduced when a good layout is used. For both common-mode and differential-mode examples using a good layout didn't completely eliminate the need for ferrites. The larger 600Ω ferrite proved to be more effective than the 100Ω ferrite, and was able to even further reduce the amount of radiated emissions from the regulators with the good layout.

To determine the overall impact that the ferrites had in reducing radiated emissions, measurement data at 70MHz was compared between all loads. The reduction in radiated emissions between the no ferrite and 600Ω ferrite configurations are plotted for both differential mode and common mode interference. Figure 3.37 and Figure 3.38 show the reduction in radiated emissions for differential mode and common mode interference respectively.

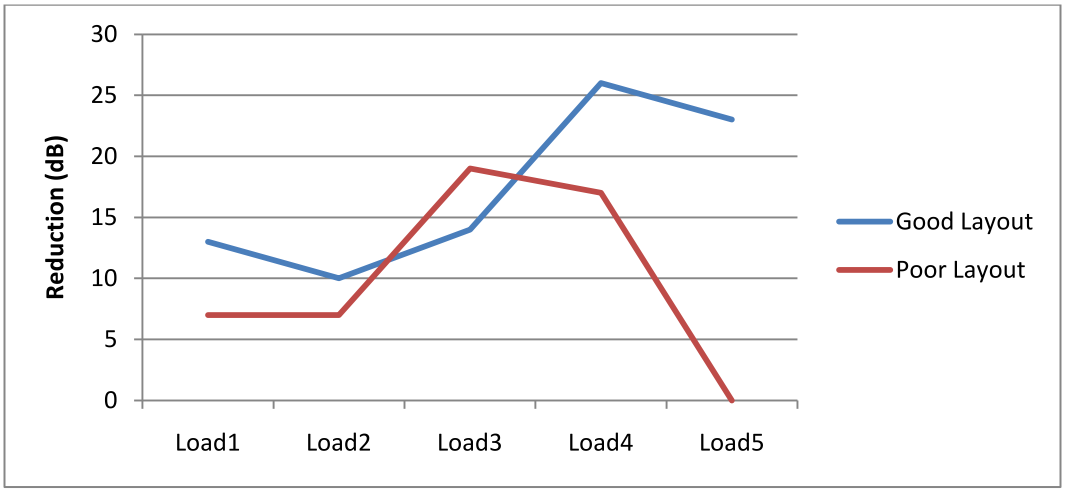


Figure 3.37. Differential Mode Reduction in Radiated Emissions.

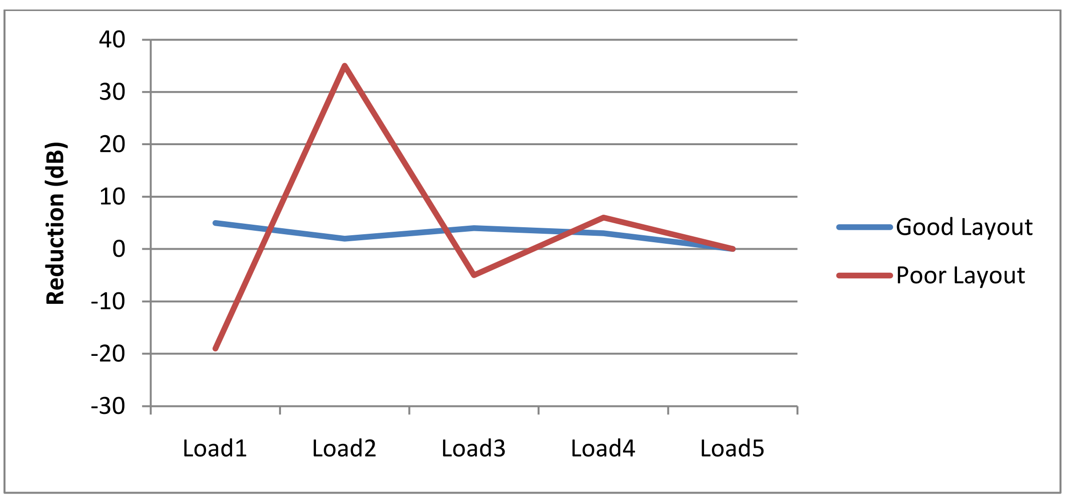


Figure 3.38. Common Mode Reduction in Radiated Emissions.

From Figure 3.37 it can be seen that load5 had the largest reduction in radiated emissions with a good layout, but with a bad layout the ferrites were unsuccessful in achieving a reduction in radiated emissions. Also from Figure 3.37, load2 and load3 had the smallest change in reduction of radiated emissions of approximately 5dB between the good and poor layout. This reiterates the importance of a good layout. The ground plane and return path under the source was successful in reducing emissions even with a poor layout. For differential mode interference, running a parallel source and return path creates the largest amount of radiated emissions and should be avoided.

From Figure 3.38, observing load1 and load 3 with a poor layout it can be seen there was a negative reduction in radiated emissions. This implies that radiated emissions actually increased with the introduction of ferrites. With a good layout, there was never an increase in radiated emissions. The increase is an indication that the circuit is having fundamental problems with its operation due to the bad layout. Load2 clearly had the largest reduction in radiated emissions between the good and poor layout of over 30dB. By forcing the load current to return under the source reduces EMI due to common mode interference by eliminating alternate paths loop currents could follow. Load5 did not have a reduction in emissions indicating that running parallel source and return paths is a bad design decision, and should be avoided.

3.2. Digital Communication

Electronic devices can communicate in either an intra- or intersystem manner. Whether it is multiple circuit boards communicating in the same system, or two entirely different systems communicating, some type of connection must be made in order to transfer data. There are multiple different communication protocols and ways to transfer data. RS232 is a very popular communication method for sending full-duplex serial data. RS232 is a form of digital communication, and therefore communicates by switching between logic high and low states, resulting in a square waveform. The quicker the slew-rate and faster the frequency of the digital communication the easier it becomes for energy to radiate away from the cable. Digital square-wave signals are made up of the superposition of several sine waves all at different frequencies. This results in a signal being transmitted which is rich in harmonic content and has the potential to radiate over a wide spectrum of frequencies.

Two computers were set up to simulate inter-system communication using RS232. An inexpensive ribbon cable was first compared to a more expensive shielded cable. Ferrites were then added to each cable to measure the reduction in radiated emissions. When establishing communication between the two computers a HyperTerminal session was started and an ASCII character “e” (0x65) was transmitted at 9600 baud. The hexadecimal value of 0x65 was chosen since there are several transitions between bits of data. As previously discussed these transitions can be a source of EMI, and therefore “e” was chosen as a worst case scenario. Successful transmission of the data was verified by observing the received character in a HyperTerminal session on the receiving computer.

3.2.1. Ribbon Cable

An inexpensive ribbon cable 1.2m long with no EMI shielding or protection was measured first. The cable was wired as a cross-over cable so that it could be connected directly between the two computers. Measurements were first made with no ferrites (Figure 3.39).

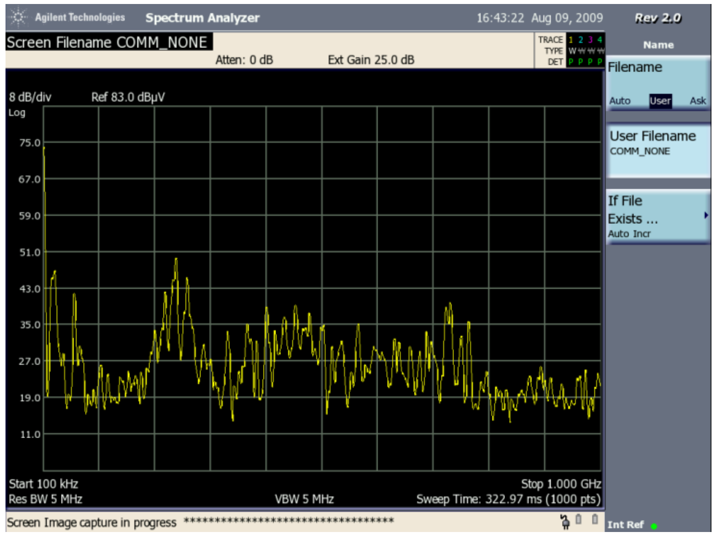


Figure 3.39. Ribbon Cable with No Ferrites.

Toroidal ferrites were added at each end of the cable, as close to each computer as possible, to reduce conducted and radiated EMI. The toroidal ferrites used are manufactured by Intermark and have a part number 212-629-3620. The toroidal ferrites are easily placed on the cable since they come in two halves which snap together with a plastic housing. Measurements were repeated with the ribbon cable and the ferrites installed (Figure 3.40).

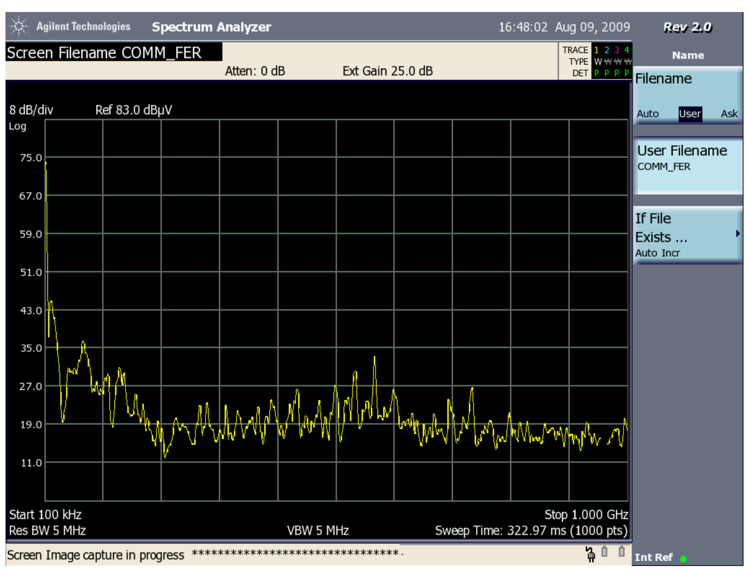


Figure 3.40. Ribbon Cable with Toroidal Ferrites.

As can be seen by adding the toroidal ferrites a significant amount of radiated EMI was reduced. Over the range of 200-800MHz a reduction of 10-30dBµV was achieved by simply adding the ferrites. A shielded cable was measured next to compare the effectiveness of the ferrites on a good quality cable.

3.2.2. Shielded Cable

Using the same measurement process as previously described, a good quality shielded cable 1.8m in length was substituted. Unlike the ribbon cable, the good quality cable has a metal braid which surrounds around all of the conductors. Each connector of the good cable is metal which makes contact to the chassis of each computer, effectively grounding the braid. With the braid grounded electric fields are terminated instead of radiating, and the overall amount EMI from the cable is reduced. The good quality cable was first measured with no ferrites (Figure 3.41).

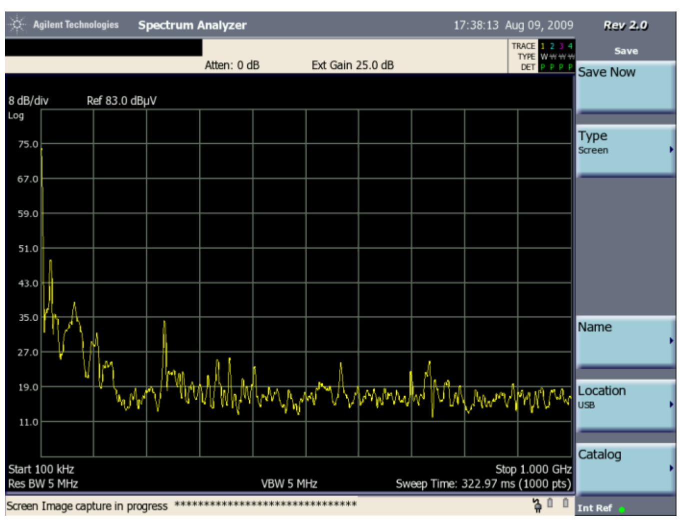


Figure 3.41. Shielded Cable with No Ferrites.

Already from Figure 3.41 it can be seen that a shielded cable with no ferrites performs very well when compared to the ribbon cable. Other than a noticeable peak at roughly 230MHz, radiated emissions are already less by approximately 5-10% than the ribbon cable with ferrites installed. Toroidal ferrites were installed on the good quality cable to see if emissions could be further reduced.

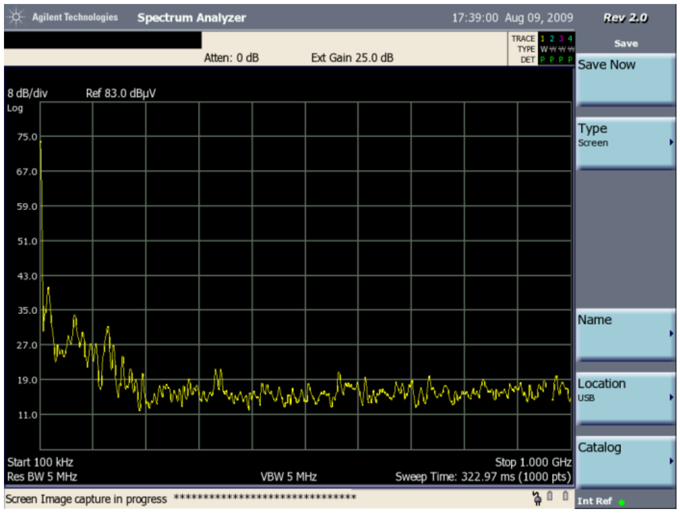


Figure 3.42. Shielded Cable with Toroidal Ferrites.

With the toroidal ferrites installed on the shielded cable (Figure 3.42) it can be seen that radiated emissions were even further reduced. Little to no EMI was measured above the noise floor above 200MHz.

The shielded cable clearly out performs the ribbon cable by decreasing radiated EMI and susceptibility. The shielded cable however costs significantly more than the cheaper ribbon cable. While designing and going through FCC testing using a shielded cable with ferrites will provide the best performance, even though it costs 4-5 times as much as the cheaper ribbon cable. Once the product has passed all FCC requirements, retesting with the lower quality cable could be done to determine if the cost is justified or necessary for the shielded cable.

3.3. Coax Cables

Coax cables are commonly used as transmission lines in high frequency devices. The geometry of the coaxial cable consists of an inner signal conductor surrounded by a braid acting as the return. The space between the signal and braid conductors is filled with a material with a high dielectric constant. Figure 3.43 shows an example of a coaxial cable.

Figure 3.43. Coaxial Cable (CommScope Properties, 2007).

As a signal is transmitted down the coaxial cable all of the energy stored in the electromagnetic field exists only between the inner signal conductor and the return braid. This geometry allows for coaxial cables to be low loss, and protects the signal against EMI.

A “leaky” cable exists when some of the electromagnetic energy from the signal is able to escape. The lost electromagnetic energy is able to radiate and act as a source of EMI. The common source for a leaky cable is the braid around the crimp on the connector has become damaged. A damaged braid around the coaxial cable will alter the return path, and change the geometry of the cable. The change in geometry will result in a change in the cable impedance, along with electromagnetic fields coupling across the break in the braid. A coax cable 0.6m in length was intentionally damaged to evaluate how much EMI was emitted due to a damaged crimp.

Three different measurements were made, the first with a perfect cable, the second with a cable that had the braid damaged and the third with a completely broken braid. The damaged braid was cut between the cable and the connector nearest to the load. Although the braid was cut it had the potential to be making intermittent contact depending on how the cable was situated, providing an inconsistent return. The cable with the completely broken braid isolated the braid and return path by 3.175mm.

The HP E4431B signal generator was used as the RF source for this experiment. The E4431B has a frequency range of 250kHz to 2GHz, which will put an upper input frequency limit on the measurements. Measurements will be made with an input frequency of 0.25GHz, 1GHz, and 1.75GHz with the signal input amplitude fixed at 100dBμV. The coaxial cable was attached to the source and terminated with a 50Ω load. The $\vec{H}$ field probe and Agilent spectrum analyzer were used as in previous experiments; however, the frequency span was increased to DC-4GHz in order to capture any harmonic content that may be present.

3.3.1. Control

A control using a good cable was measured first. The control had a good crimp and no damage to the cable. Measurements were made at the source, at the load, and the middle of the cable. Measurements were taken at the same frequency of the input signal and at the three different locations to determine the impact at higher frequencies (Figures 3.44, 3.45 and 3.46).

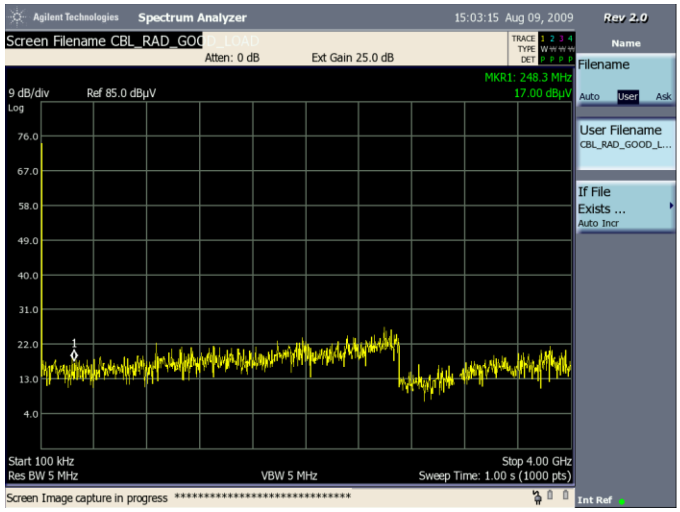


Figure 3.44. Good Cable – Radiated at Load (Input Signal Frequency = 0.25GHz).

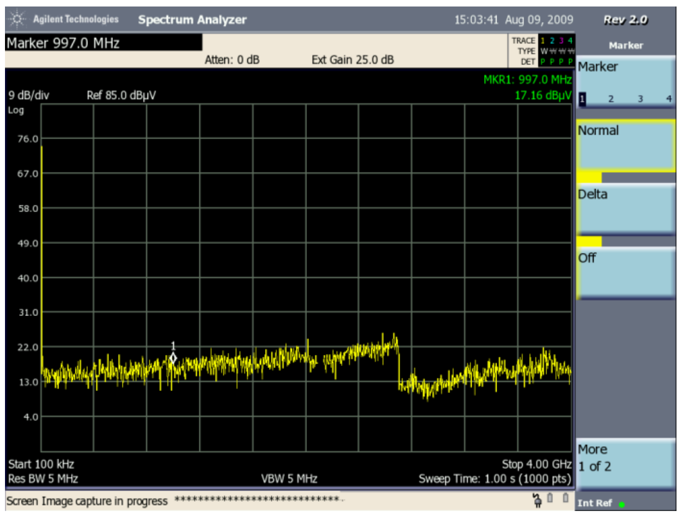


Figure 3.45. Good Cable – Radiated at Load (Input Signal Frequency = 1.0GHz).

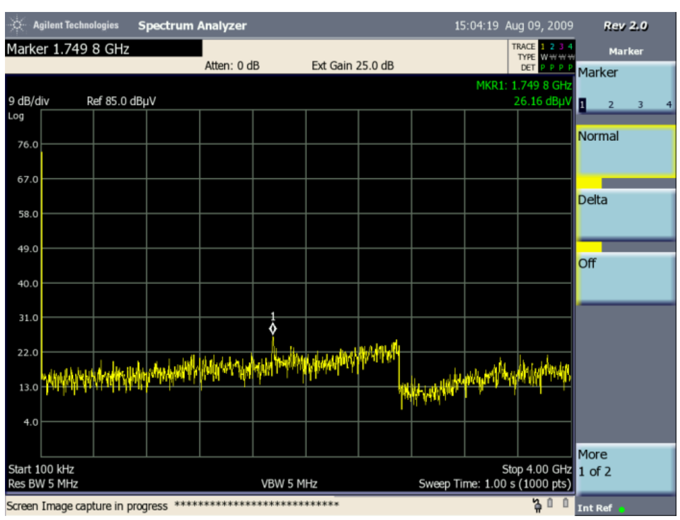


Figure 3.46. Good Cable – Radiated at Load (Input Signal Frequency = 1.75GHz).

The marker tool on the spectrum analyzer was used to take measurements at the same frequency as the input. From Figure 3.44 through Figure 3.46, it can be seen that very little EMI is leaked from a good cable. Only at 1.75GHz does it become apparent that some electromagnetic energy is starting to escape from the cable. A large span was captured to determine if any harmonic content is present. It is noticeable that a ‘discontinuity’ is present at roughly 2.7GHz. This is an effect of the Agilent spectrum analyzer in wide-band frequency measurements. Figures for measurements at the source and middle locations are consistent with the results from Figure 3.44 through Figure 3.46, and only show the noise floor. The data for the source and the middle measurements can be found in the appendix.

3.3.2. Damaged Braid

A cable was intentionally damaged to simulate a bad crimp. The return braid was cut around the connector on the cable nearest to the load. Although the braid was cut it was still touching providing an inconsistent return.

Observing measurements at the load (Figure 3.47 through Figure 3.49), a significant amount of radiated emissions are present. This is due to the geometry of the coaxial cable being changed by the braid discontinuity, and electromagnetic fields coupling across the return braid to the connector. As the frequency increases, the amount of radiated emissions also increases.

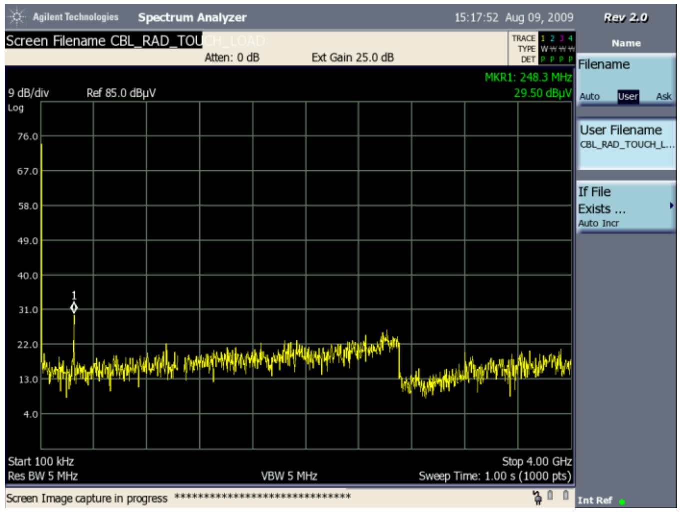


Figure 3.47. Damaged Cable – Radiated at Load (Input Signal Frequency = 0.25GHz).

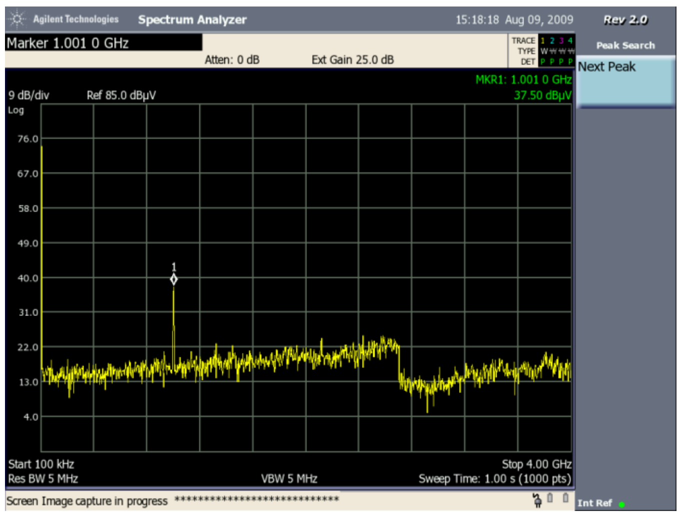


Figure 3.48. Damaged Cable – Radiated at Load (Input Signal Frequency = 1.0GHz).

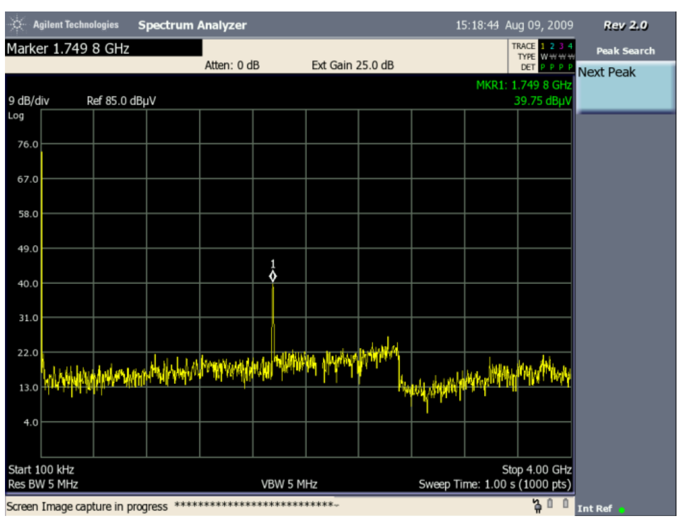


Figure 3.49. Damaged Cable – Radiated at Load (Input Signal Frequency = 1.75GHz).

The cut in the coax return braid changes the cables return path, along with its impedance and causes reflections. The reflections maximize the current on the cable and increase the amount of radiated emissions. Radiated emissions are observed at the generator with the damaged braid as seen in Figure 3.50 through Figure 3.52.

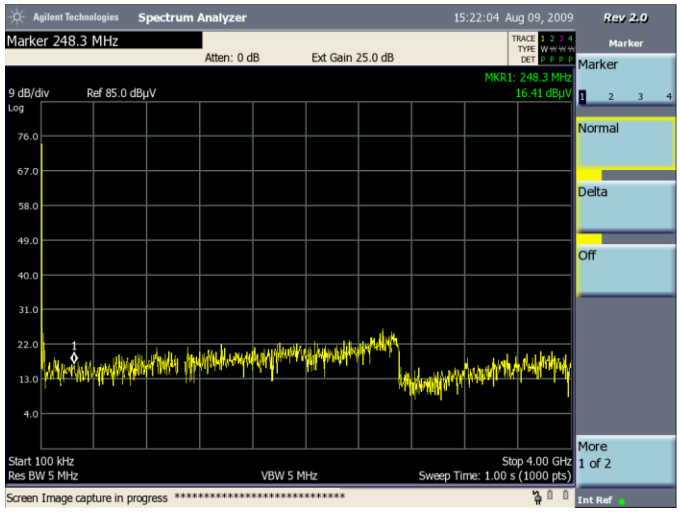


Figure 3.50. Damaged Cable – Radiated at Generator (Input Signal Frequency = 0.25GHz).

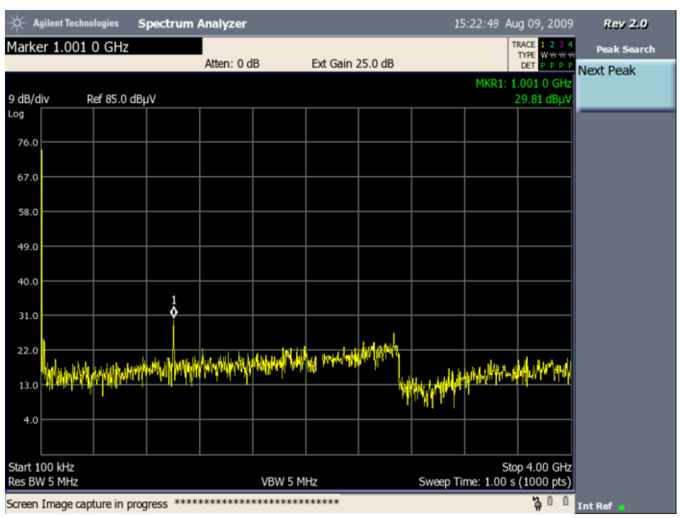


Figure 3.51. Damaged Cable – Radiated at Generator (Input Signal Frequency = 1.0GHz).

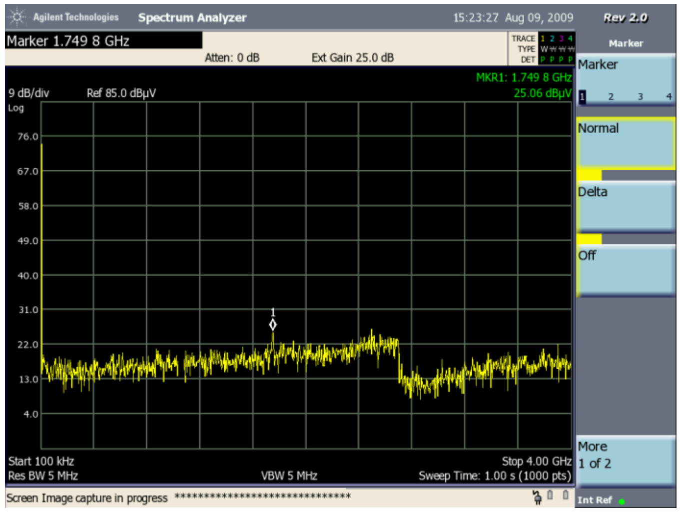


Figure 3.52. Damaged Cable – Radiated at Generator (Input Signal Frequency = 1.75GHz).

Although the amount of radiated emissions is significantly less at the source than the load some EMI is still radiating. The impedance mismatch created from the damaged braid at the load is reflecting energy back to the source, which is then radiating at the source connector. However, only higher frequencies are radiating. At 0.25GHz there is no radiated emission at the source. Measurements for the middle of the cable only show the noise floor, and can be found in the appendix.

3.3.3. Broken Braid

The damaged braid was completely isolated from the connector by 3.175mm to remove any chance of an intermittent connection. This was done to simulate a crimp that has completely failed. The measurement process was performed as previously.

The large break in the return braid allowed an excessive amount of electromagnetic energy to escape. Following the same logic as the damaged cable the change in geometry has a significant impact on the return path and impedance of the cable. The broken return path maximizes current on the cable, and maximizes radiated emissions. Some energy is also escaping as the electromagnetic field couples across the gap. A significant amount of radiated emissions were radiated from the source and middle of the cable. Measurement data at the generator and middle for the broken cable are consistent with the load data and can be found in the appendix. Figure 3.53 through Figure 3.55 show the broken cable radiated emissions at the load.

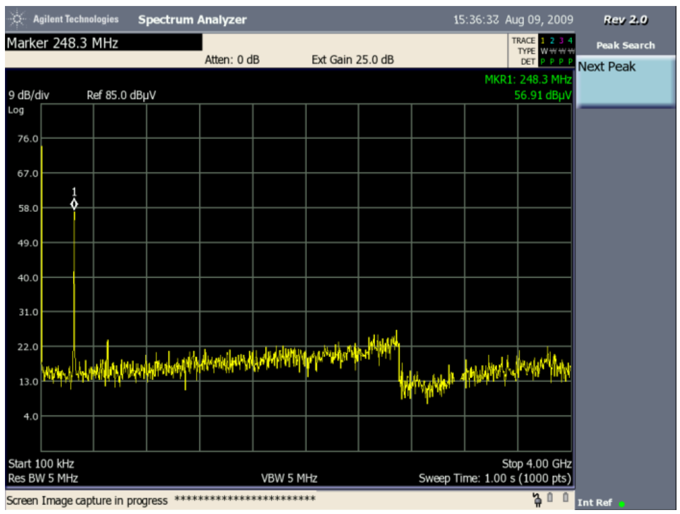


Figure 3.53. Broken Cable – Radiated at Load (Input Signal Frequency = 0.25GHz).

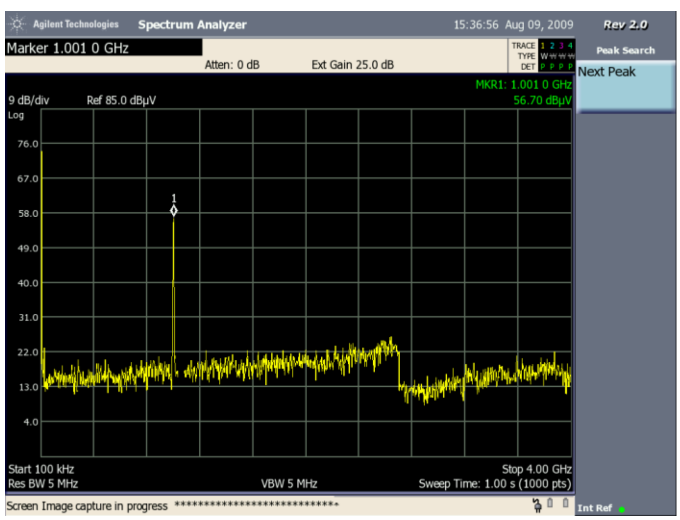


Figure 3.54. Broken Cable – Radiated at Load (Input Signal Frequency = 1.0GHz).

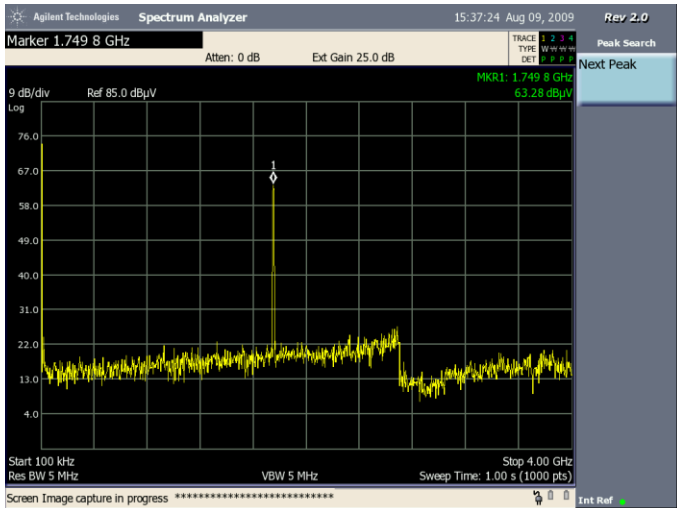


Figure 3.55. Broken Cable – Radiated at Load (Input Signal Frequency = 1.75GHz).

When evaluating the data, it can be seen that the good cable had little radiated EMI, only becoming noticeable at higher frequencies. The cable with the damaged braid had radiated emissions at both the load and the source. The cable with the completely broken braid radiated everywhere on the cable. Changing the cable geometry has a significant impact on the cable impedance. As the gap between the braid and connector becomes larger, the change in impedance increases. The larger the impedance difference becomes, the more radiated EMI and losses within the cable can be expected.

3.4. Low Noise Amplifiers

Low noise amplifiers (LNAs) have the potential to cause multiple EMI related problems. Since LNAs typically operate in the RF spectrum, it is much easier for electromagnetic energy to radiate into free space or couple into nearby signal paths. An LNA also tends to be mostly linear in its operating range, however as the LNA becomes saturated and begins to compress it becomes non-linear. These non-linear effects will result in cross-products and other harmonics which will become conducted EMI and have the potential to radiate.

Due to several EMI related issues that can arise from using LNAs, multiple circuits were designed to exacerbate several of these issues. The LNA chosen for this experiment was the MAX2640 Ultra-Low Noise LNA. This LNA was chosen due to its small package size, low external component count, wide frequency capabilities, and ultra-low noise characteristics. The LNA output is designed for 50Ω systems, a matching network must be used on the input to match to 50Ω. Circuits designed include a control, a LNA with a load impedance mismatch, a LNA with a monopole antenna attached to it and a LNA with a parallel micro-strip transmission line.

3.4.1. Control

The control LNA was designed to specifications outlined by the manufacturer to yield the best results. These specifications include input and output network matching component values and layout gerber plots showing ideal component placement. Figure 3.56 shows the LNA schematic for the control LNA.

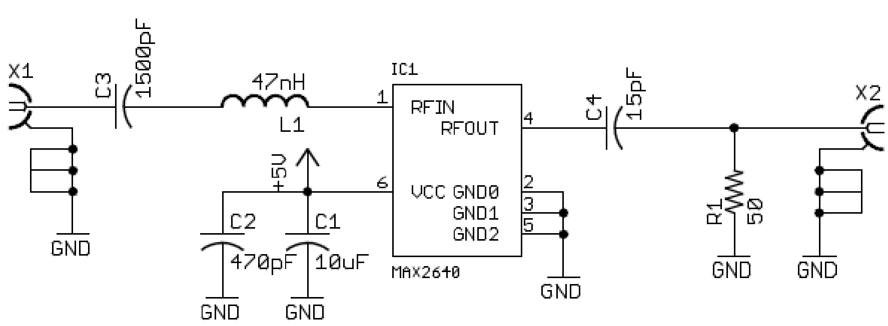


Figure 3.56. Control LNA Schematic.

From the schematic (Figure 3.56), connectors X1 and X2 are SMA connectors for measuring the gain of the LNA. The resistor R1 was not populated when measuring the gain. R1 is later populated to terminate the control LNA when taking radiated measurements. Components C3 and L1 are used for input impedance matching, and C4 is an output DC blocking capacitor. Figure 3.57 shows the PCB layout of the Control LNA.

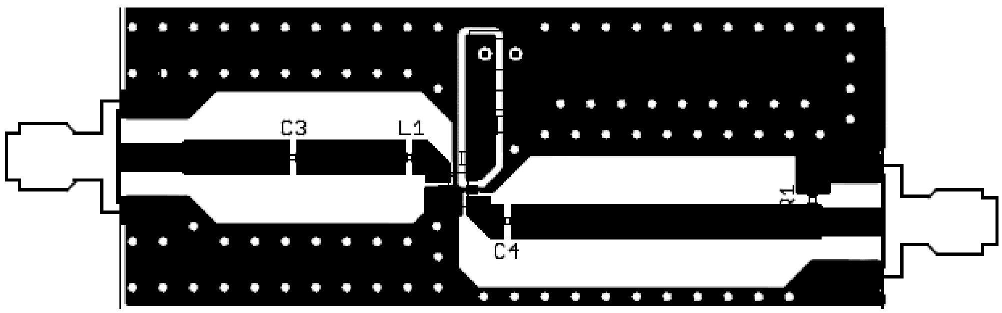


Figure 3.57. Control LNA Layout.

The HP E4431B signal generator, as used in previous experiments, was used as the RF source for the LNA. The control LNA was measured and found to have a gain of 15.7dB. After verifying proper circuit operation, the following conditions outlined in Table 3.2 were used to measured radiated EMI from the LNA while in normal operating mode, and in compression.

Table 3.2.
LNA Input Parameters.

	Frequency (MHz)	Input Power Level (dBm)
Normal Operation	434	-50
Compression	434	-10

Using the input parameters from Table 3.2, radiated EMI measurements were made of the control LNA for normal operating mode and in compression. All measurements were captured using the Agilent spectrum analyzer as in previous experiments; however, the span was changed from DC-2.5GHz to measure harmonics and cross-products. In previous experiments the external gain on the spectrum analyzer was set to 25dB to compensate for the HP 8447D pre-amp. While taking all LNA measurements the pre-amp was used; however, the spectrum analyzer did not have the external gain compensation set to 25dB. Therefore all LNA measurements are relevant to one another, but measured radiated emissions are 25dB larger than actual. Measurements were made using the probe placed directly above the MAX2640. Figures 3.58 and 3.59 show the control LNA when operation within normal input parameters, and in compression.

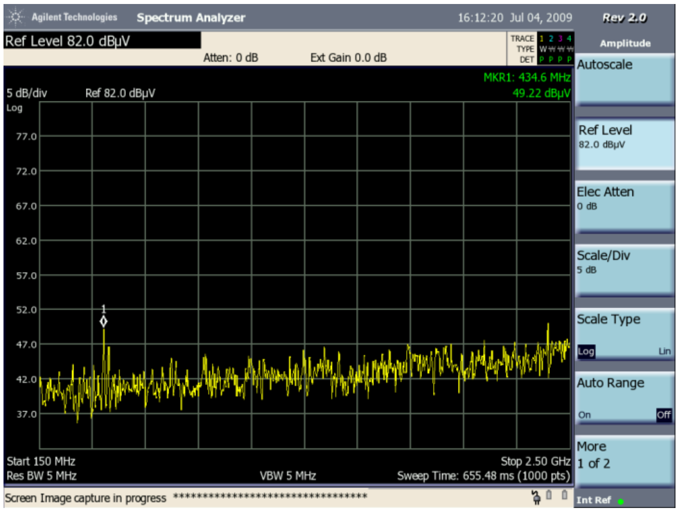


Figure 3.58. Control Normal Operating Conditions.

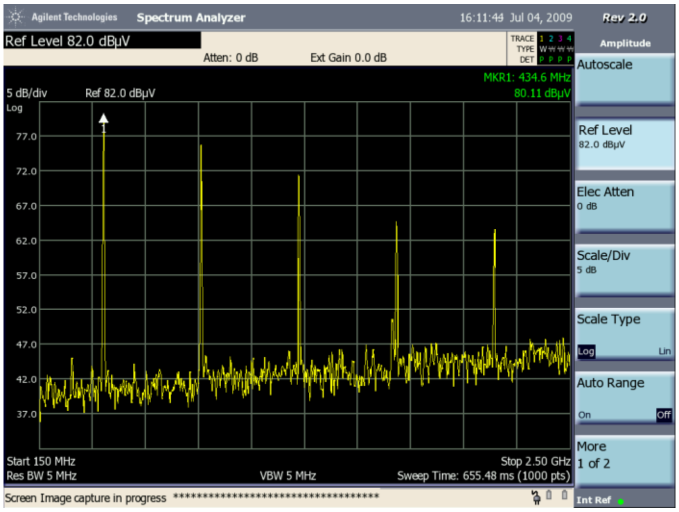


Figure 3.59. Control in Compression.

The control LNA in normal operating mode is quiet except at the fundamental operating frequency. However, under compression with a much larger input power the amount of radiated EMI increases drastically. It is understood that with an increase of input power the amount of radiated emissions will also increase. For a 40dB increase of input power there was a 30dBµV increase in radiated emissions. The second through fifth harmonics also become present and radiate a significant amount. Table 3.3 shows the measured radiated emissions from the control LNA.

Table 3.3.
Control LNA Radiated Measurements.

	Frequency (MHz)	Radiated Emissions (dBµV)
Normal Operation	434	49.22
Compression	434	80.11

The measured values from Table 3.3 will be compared to the incorrectly designed LNAs to determine the percent difference in radiated emissions. It is already clear that operating a correctly designed LNA in compression can cause unwanted EMI related problems. The control LNA will now be compared to LNAs which also have design related issues.

3.4.2. Impedance Mismatch

The LNA with an impedance match is identical to the control except for the load is no longer matched. From Figure 3.56 the load resistor (R1) was replaced with a 100Ω resistor instead of 50Ω. The idea was to maximize reflections, increase the current on the output transmission line, and increase radiated emissions. Measurements were made using the input parameters from Table 3.2. Observing Figure 3.60 and Figure 3.61 it can be seen that the unmatched LNA is radiating slightly more than the control LNA.

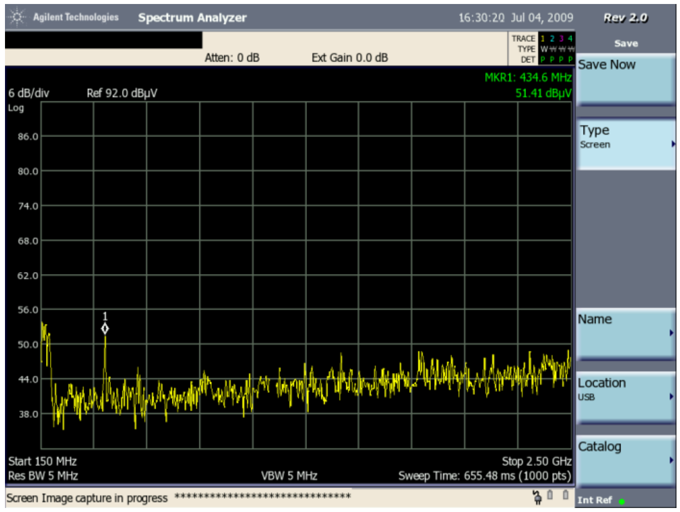


Figure 3.60. Unmatched Normal Operating Conditions.

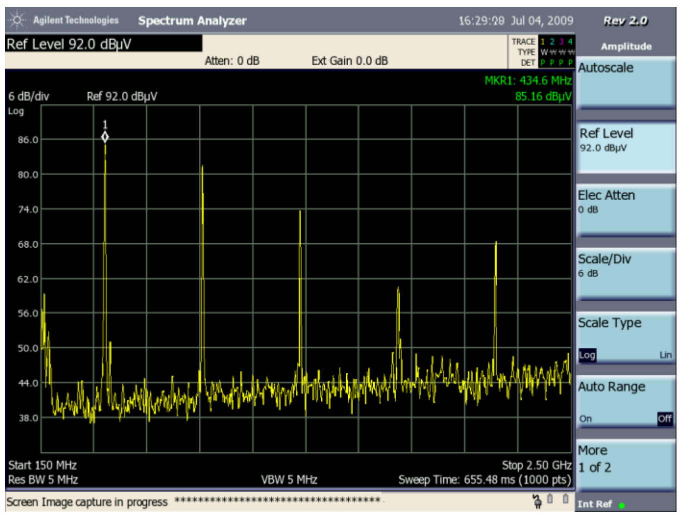


Figure 3.61. Unmatched LNA in Compression.

Table 3.4 shows the percent change in radiated emissions of the unmatched LNA when compared to the control.

Table 3.4.
Unmatched LNA Radiated Measurements.

	Frequency (MHz)	Control Radiated Emissions (dBµV)	Unmatched Radiated Emissions (dBµV)	Percent Change (%)
Normal Operation	434	49.22	51.41	4.45
Compression	434	80.11	85.16	6.30

From Table 3.4 there is an increase in radiated emissions over the control for both normal operation and in compression. The load mismatch is causing reflections at the output and increasing the amount of EMI. The percent change when operating in normal operation and compression is also not constant. This shows that the two measurements are not proportional, and when operating in compression there is an increase in radiated EMI. Comparing Figure 3.59 to Figure 3.61, it can be seen that the second harmonic of the unmatched LNA is radiating approximately the same amount as fundamental of the control. This shows an increase in radiated emissions over all the harmonic content as well, and not just at the fundamental.

3.4.3. Antenna Effects

Design flaws can cause some PCB traces to act as antennae. The control LNA can be easily modified to create an antenna. Removing the load resistor, increasing the length of the output micro-strip line, and removing the ground plane under the output transmission line forms a monopole antenna. By intentionally creating an antenna, it would be expected that the amount of radiated emissions would increase. Figure 3.62 shows the schematic of the LNA with the antenna.

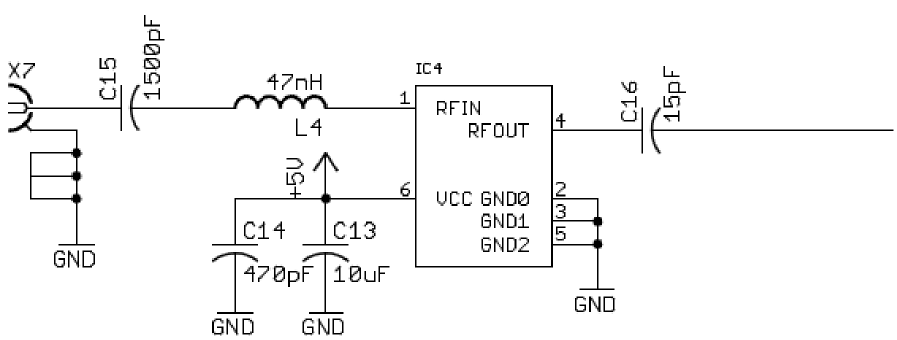


Figure 3.62. Antenna LNA Schematic.

From Figure 3.62, the load resistor from the control has been removed, the micro-strip line after the DC blocking capacitor (C16) was increased to a quarter-lambda in length. The quarter-lambda length was chosen since monopole antennas are multiples of this length. A quarter-lambda on FR4 substrate is calculated as follows:

$$\frac{\lambda}{4} = \frac{c}{f\sqrt{\varepsilon_r}} \tag{3.1}$$

Where c is the speed of light, f is the frequency, and ε_r is the relative dielectric constant of the substrate. Substituting the necessary values into equation 3.1 a quarter-lambda can be calculated to be:

$$\frac{\lambda}{4} = \frac{3x10^8}{434MHz\sqrt{4.8}} = 78.87mm \tag{3.2}$$

Increasing the length of the output micro-strip line to 78.87mm will yield a quarter-wavelength antenna on the output of the LNA. Figure 3.63 shows the layout of the LNA with the attached monopole antenna.

Figure 3.63. Antenna LNA Layout.

To maintain consistency with previous measurements made of the control LNA and the impedance mismatched LNA, measurements were again made with the probe placed directly over the MAX2640. The input parameters for the LNA are described in Table 3.2. As expected, the LNA with the monopole antenna radiated a significant amount more than the control. From Figure 3.64 and Figure 3.65 it can be seen that there was a very large increase in radiated emissions, even operating within normal operating conditions.

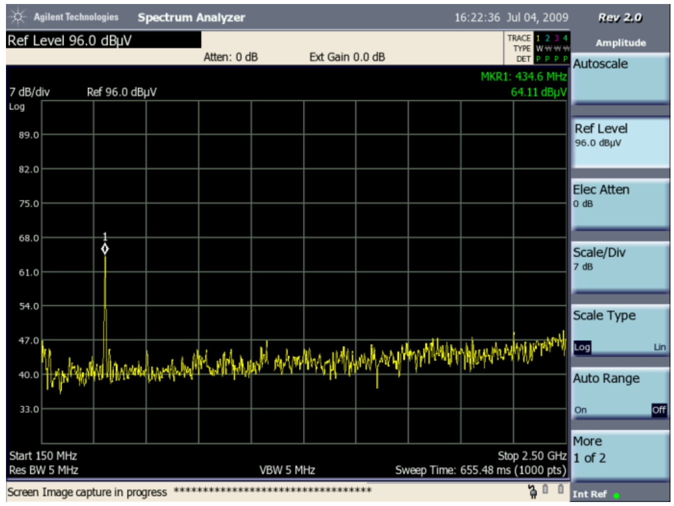


Figure 3.64. Antenna Effects Normal Operating Conditions.

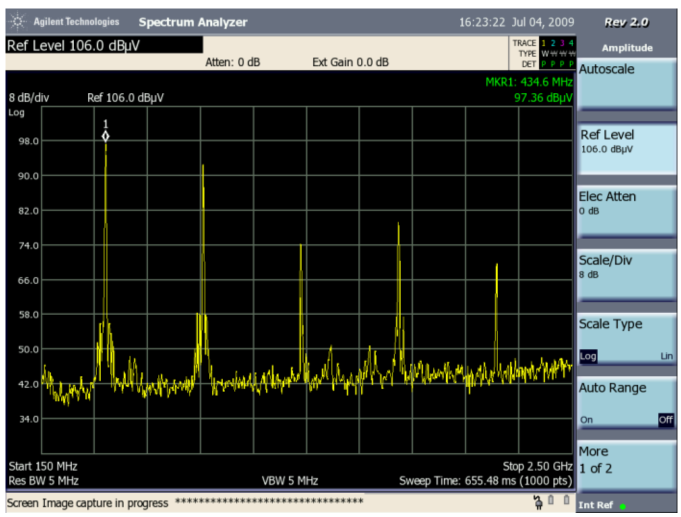


Figure 3.65. Antenna Effects in Compression.

Table 3.5 shows the percent change in radiated emissions of the antenna LNA when compared to the control.

Table 3.5.
Antenna LNA Radiated Measurements.

	Frequency (MHz)	Control Radiated Emissions (dBμV)	Antenna Radiated Emissions (dBμV)	Percent Change (%)
Normal Operation	434	49.22	64.11	30.25
Compression	434	80.11	97.36	21.53

From Table 3.5 it can be seen that there is a 30.25% increase in radiated emissions when in normal operation, and 21.53% increase when in compression. This is a large increase in the amount of radiated EMI, and shows the importance of using proper PCB layout techniques such as keeping trace lengths as short as possible and using ground planes for return currents. It is interesting to note that when in compression, there was a smaller percent increase in radiated emissions than normal operation. This may be caused by a loading issue. Since there is no load resistor on the output multiple reflections on the line may cause destructive interference, and actually reduce the amount of radiated emissions.

3.4.4. Coupling

Two signals running parallel to one another have the potential to couple into one another. In RF layouts, there is a general 2*w* rule of keeping nearby signals or ground planes twice the width (*w*) of the signal trace away. Following this rule keeps interference and losses due to coupling at a minimum. The output microstrip line from the LNA was extended to a quarter-wavelength and terminated into a 50Ω resistor. A parallel microstrip line was run next to the signal line for the EM fields to couple into, and a SMA connector was placed on the output of the parallel microstrip line. Figure 3.66 shows the coupling LNA schematic, and Figure 3.67 shows the layout.

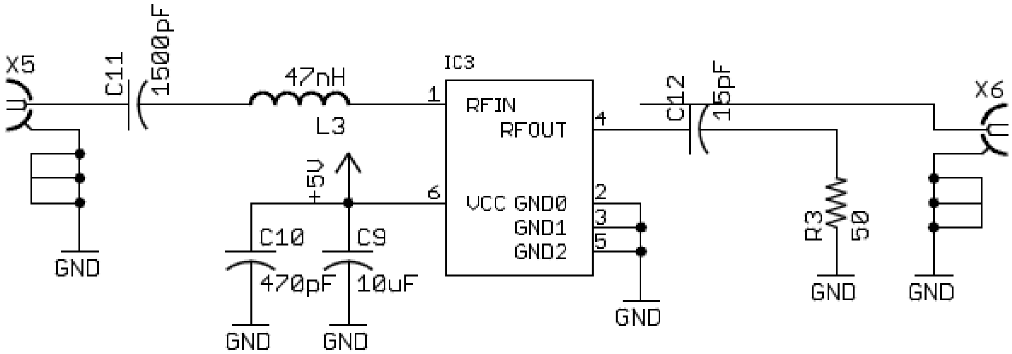


Figure 3.66. Coupling LNA Schematic.

Figure 3.67. Coupling LNA Layout.

From Figure 3.67 it can be seen that the 2*w* rule is not being followed. As a function of *w*, the distance between the two micro-strip lines is 0.02*w*. Effectively, the distance between the two micro-strip lines are 100 times closer than a good design.

The conducted EMI was measured using the spectrum analyzer without the pre-amplifier. Unlike the radiated measurements made with the previous LNAs, the coupling LNA is measuring the conducted EMI. The SMA connector on the isolated micro-strip line is connected directly to the input of the spectrum analyzer. The input parameters described in Table 3.2 are used when setting up this experiment. When observing the results from Figures 3.68 and 3.69, the amount of energy which is coupling into the parallel trace is of concern.

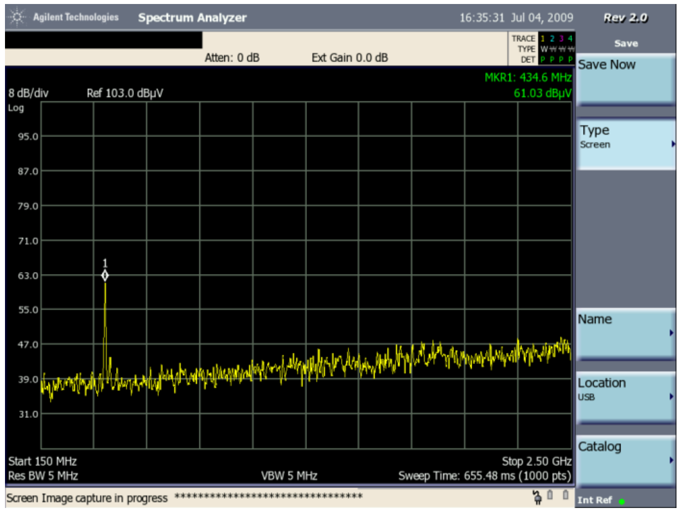


Figure 3.68. Conducted EMI Normal Operation.

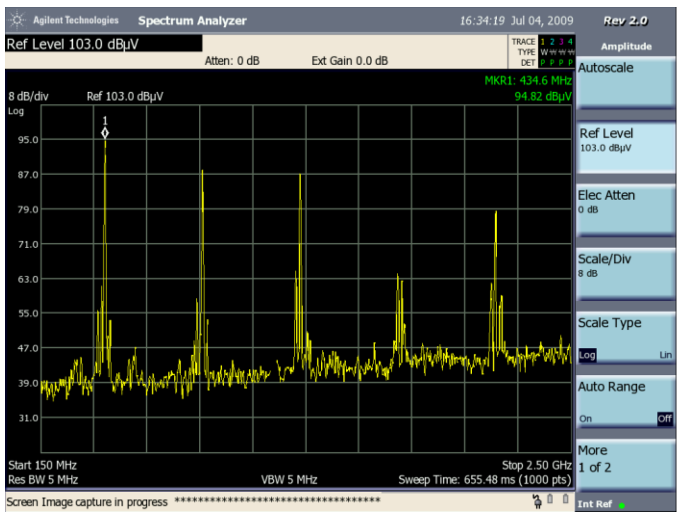


Figure 3.69. Conducted EMI in Compression.

From these results, it can be seen that a large amount of energy is able to couple into the parallel microstrip. In a 50Ω system the conversion between dBm and dBµV is:

$$dBm = dB\mu V - 107 \tag{3.3}$$

In normal operating mode, the input is -50dBm and the LNA has a gain of 15.7. Using these values the output can be calculated to be 72.7dBµV. From Figure 3.68, the measured output from the coupled signal is 61.03dBµV. This shows a very large amount of energy was able to couple into the nearby trace. Had this been implemented in an actual system this energy would propagate throughout the system forming complex loops which are almost certain to radiate. When the LNA is in compression, the gain is no longer 15.7dB which complicates the calculation, but since a significant amount of coupling takes effect in normal operation an equal if not larger amount will also occur while in compression. It can be seen from Figure 3.69 that harmonics and cross products also couple into the nearby microstrip line.

To compare radiated emissions between the control, unmatched, and antenna effects LNAs, a plot was created showing the overall radiated emissions at the fundamental frequency. Figure 3.70 shows the overall radiated emissions from each LNA.

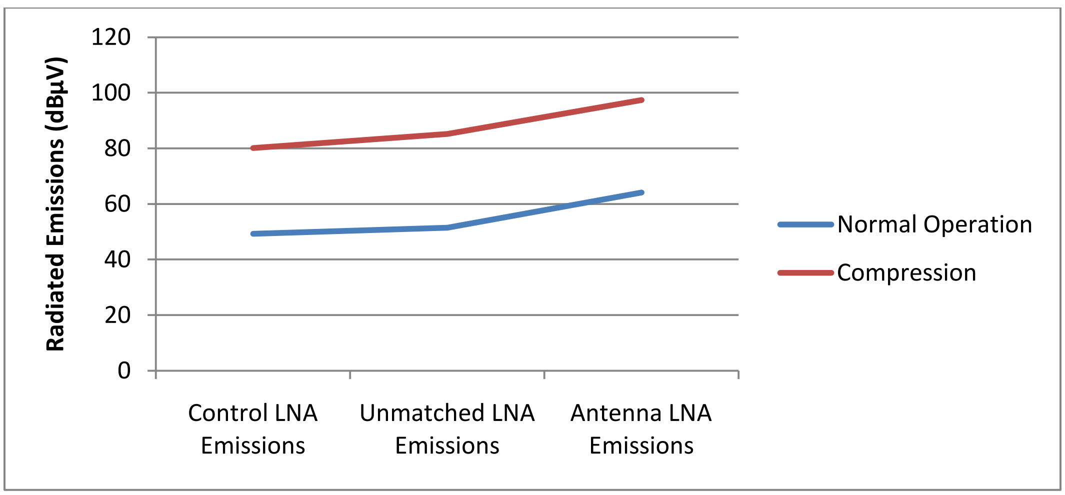


Figure 3.70. LNA Radiated Emissions Comparison.

From Figure 3.70 it can be seen that at the fundamental frequency that the overall increase in radiated emissions for normal operation and compression is approximately 35dB higher for all three LNAs. It needs to be noted that when operating in compression EMI due to harmonics will also be present in the system, where in normal operation the effects of the harmonic content is negligible. As shown in Figure 3.70, the control LNA has the lowest amount of radiated emissions at 49dB, the unmatched LNA radiated emissions is approximately 2dB higher than the control due to reflections, and the antenna LNA radiated emissions is 15dB higher than the control demonstrating the antenna is having a significant impact. From these results it can be concluded that unintentional antennas on a PCB can drastically increase the amount of radiated EMI. Unintentional antennas will result in higher radiated emissions than unmatched impedances.

3.5. Case Study: Mixed Analog and Digital Circuit Board

The previous work attempted to isolate individual circuits which if designed or operated incorrectly could contribute to conducted and radiated EMI. In a typical product many of the previous circuits could be used all in the same design. The combined radiation from multiple circuits could be significant enough that a product may fail FCC certification. As discussed previously, a product which fails FCC may require significant design changes which could add to the overall development cost of a product.

A case study is presented on a circuit board used in a product which incorporates many of the individual designs previously evaluated. This circuit board uses switching regulators for DC-DC power conversion, multiple RS232 connections are used to communicate with peripherals on and off the board, coaxial cables are used to carry signals susceptible to EMI, and multiple sensitive analog and digital signals are present on the board. Careful design considerations were implemented in the board during the initial layout in order to reduce conducted and radiated emissions.

The board has already been optimized to minimize radiated EMI following proper layout techniques discussed in the previous chapter. This includes proper board zoning, special consideration when routing critical traces, ground planes, filters, and ferrites. Since much of the development has been completed for this product, very little of the EMI techniques used in the design can be changed. One thing that can be changed is the size of the ferrites included in the design. Measurements were made in order to make conclusions about the overall effectiveness of the ferrites used in the design.

Due to the size of the board it was partitioned into six separate sections. Of the different sections, each had a specific purpose in the operation of the board. Table 3.6 breaks down each section and details the circuit operation.

Table 3.6.
Case Study Board Partitions.

PCB Section	*Circuit Operation*
1	Switching Regulators
2	Digital Information Processing
3	Digital to Analog Conversion
4	Analog Format Conversion
5	Analog to Digital Conversion
6	Digital Communication

Each section was measured with the board in both normal operating mode and in standby. When the board is in standby most of the high frequency switching components were held in reset. The board was first measured with no ferrites.

From the data (Figure 3.71 through Figure 3.82) it can be seen that the different sections of the PCB are all radiating uniquely. When the PCB is held in standby, the overall noise floor drops, allowing for identification of individual frequency components radiating the most. Individual frequency components can be clearly identified from Figure 3.74. Section 3 through section 5 (Figure 3.75 through Figure 3.80) are affected the most by placing the board in standby. Radiated emissions from section 5 decrease by roughly 15-20dBμV by placing the components in standby.

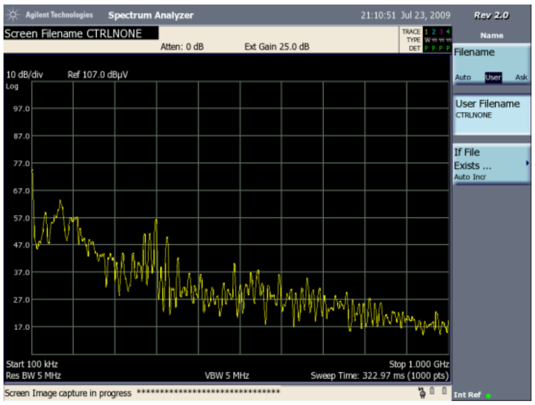


Figure 3.71. Section 1 No Ferrites.

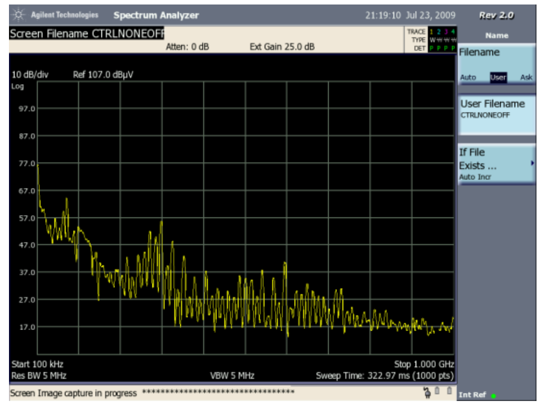


Figure 3.72. Section 1 No Ferrites Standby.

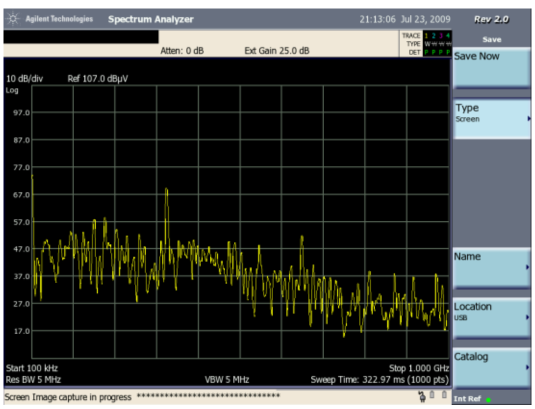


Figure 3.73. Section 2 No Ferrites.

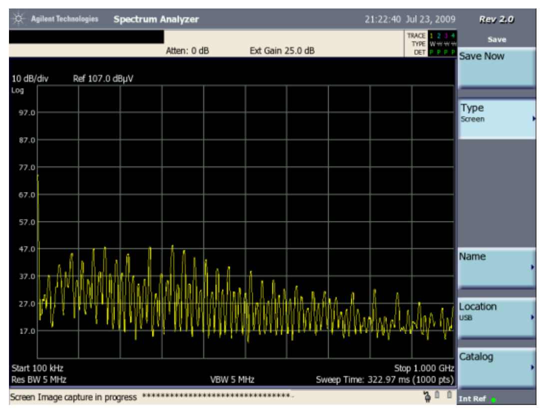


Figure 3.74. Section 2 No Ferrites Standby.

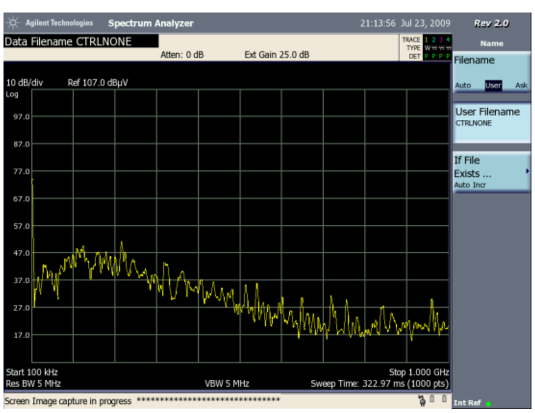


Figure 3.75. Section 3 No Ferrites.

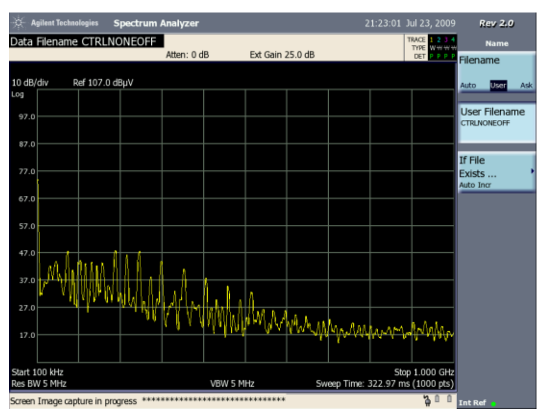


Figure 3.76. Section 3 No Ferrites Standby.

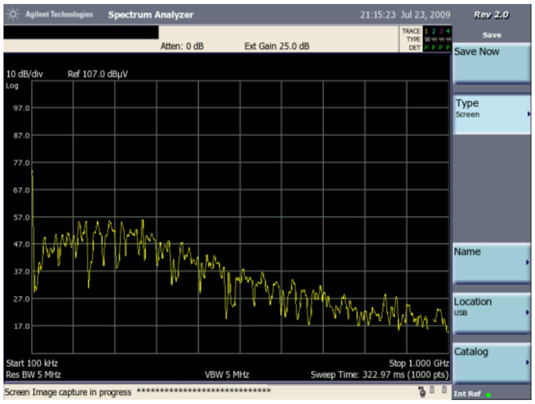


Figure 3.77. Section 4 No Ferrites.

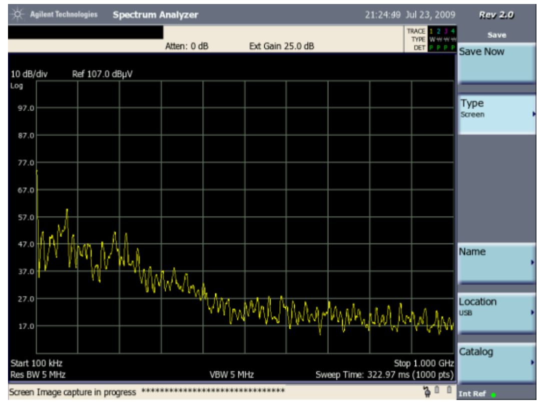


Figure 3.78. Section 4 No Ferrites Standby.

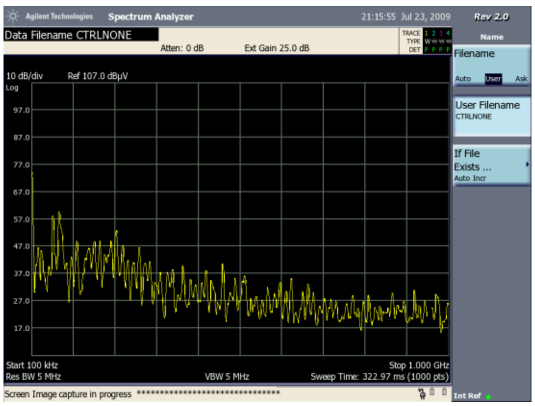


Figure 3.79. Section 5 No Ferrites.

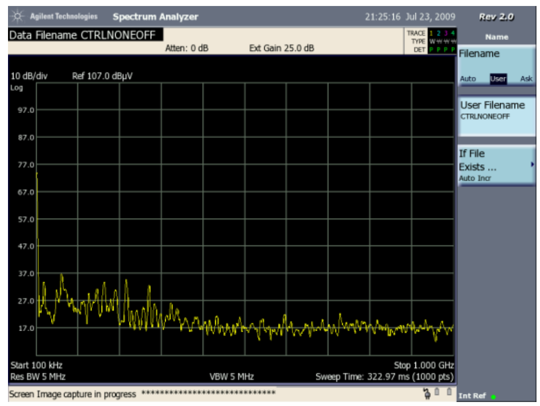


Figure 3.80. Section 5 No Ferrites Standby.

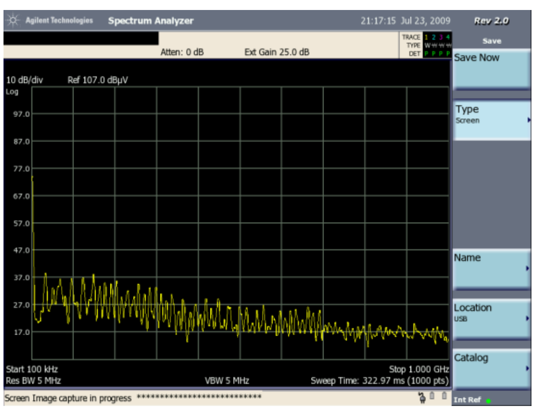


Figure 3.81. Section 6 No Ferrites.

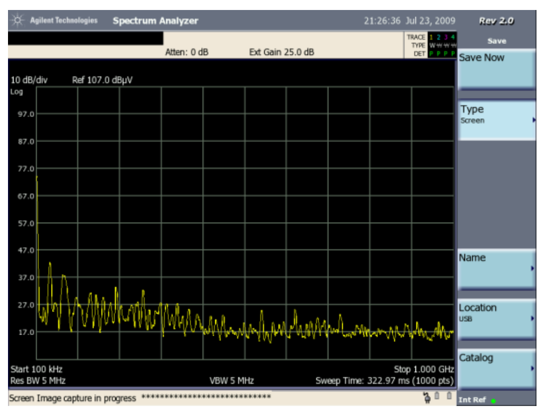


Figure 3.82. Section 6 No Ferrites Standby.

Ferrites were installed on only the digital portion of circuits on the PCB next. All ferrites installed were 30Ω 1A surface mount bead ferrites. Figure 3.83 through Figure 3.94 show the radiated EMI with ferrites added on the digital circuitry. Observing the results with the digital ferrites installed there is little to no difference in measured results. This could be due to a few issues. The first being that the analog section is radiating excessively over the digital portion, second being that the 30Ω 1A ferrites are simply not large enough, and the third being the PCB is already optimized.

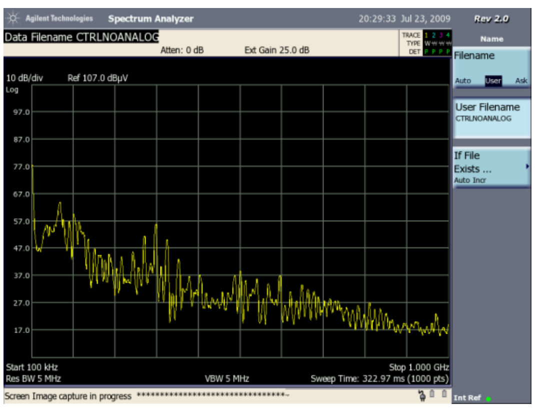


Figure 3.83. Section 1 Digital Only.

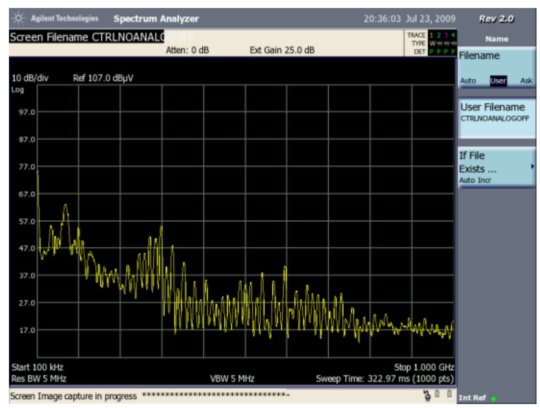


Figure 3.84. Section 1 Digital Only Standby.

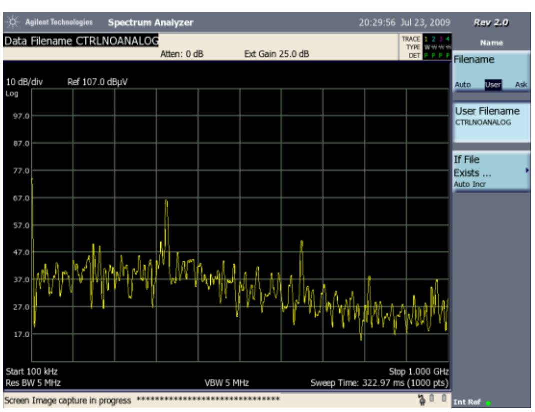


Figure 3.85. Section 2 Digital Only.

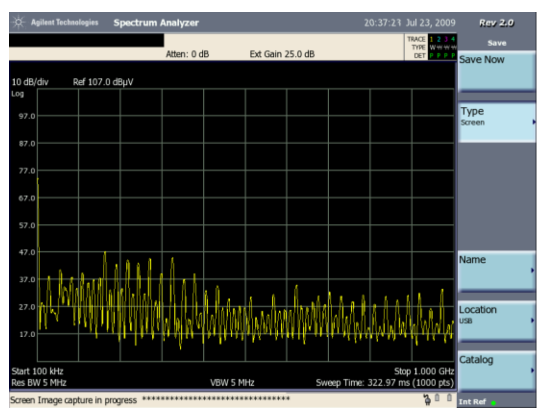


Figure 3.86. Section 2 Digital Only Standby.

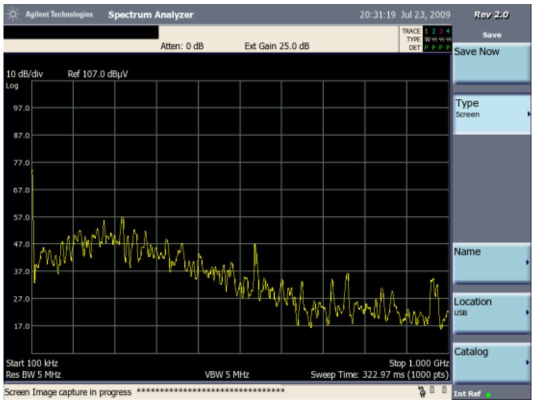


Figure 3.87. Section 3 Digital Only.

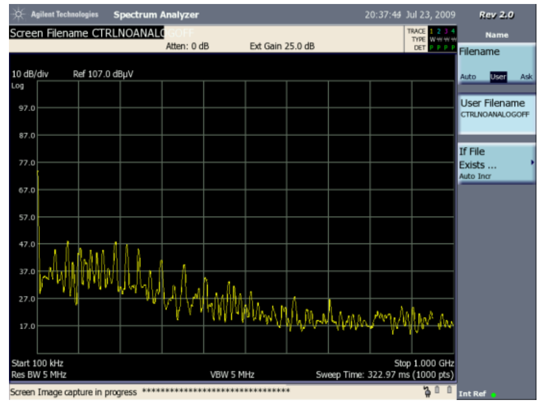


Figure 3.88. Section 3 Digital Only Standby.

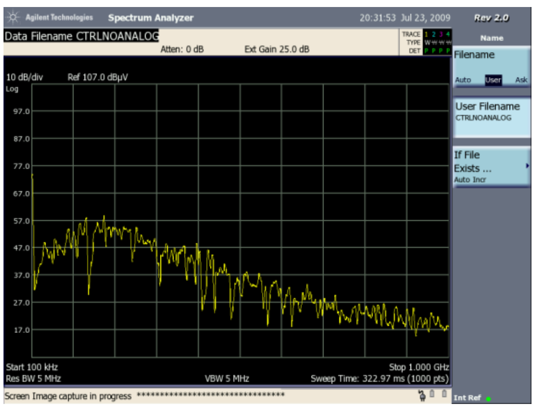


Figure 3.89. Section 4 Digital Only.

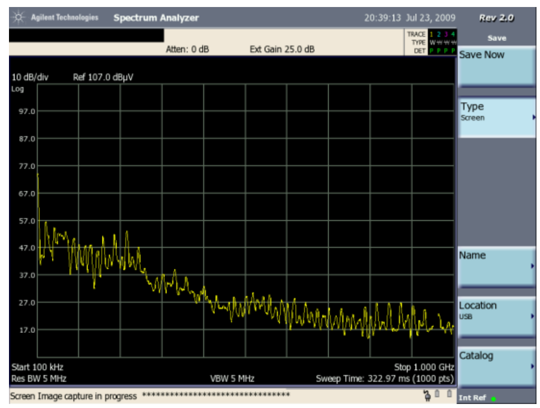


Figure 3.90. Section 4 Digital Only Standby.

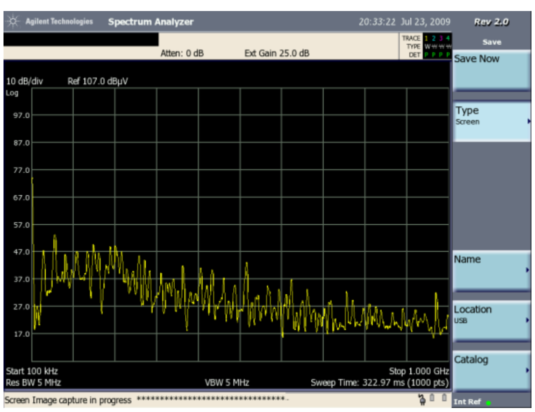


Figure 3.91. Section 5 Digital Only.

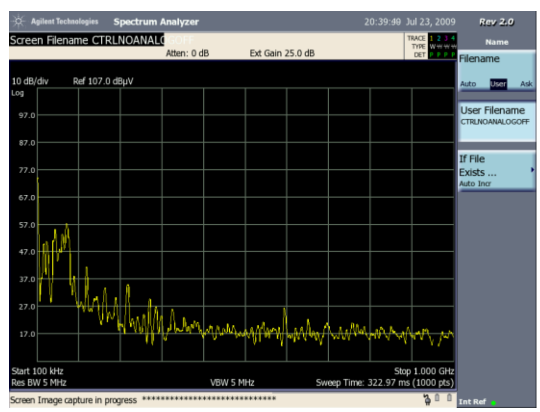


Figure 3.92. Section 5 Digital Only Standby.

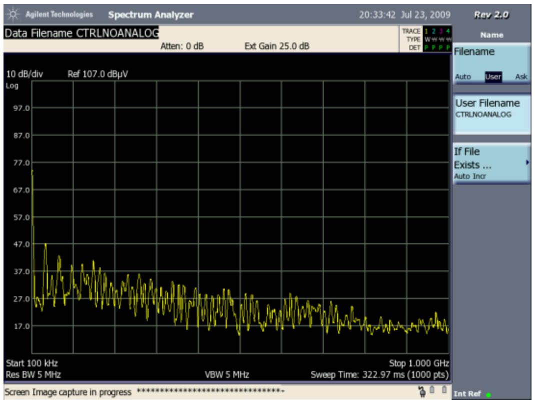


Figure 3.93. Section 6 Digital Only.

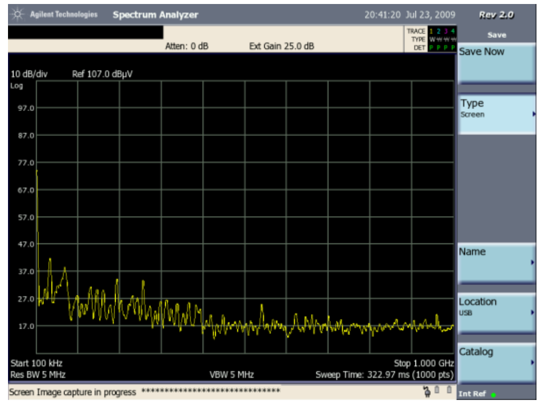


Figure 3.94. Section 6 Digital Only Standby.

Next, the ferrites for the analog section were added to the board. With ferrites installed for both the analog and digital sections ideally this setup should provide the lowest amount of radiated EMI. With ferrites installed on both the analog and digital sections of the PCB a slight improvement can be observed in many of the measurements over the captures with no ferrites. The largest improvement can be observed in Section 5 (Figure 3.103) with a reduction of 10dB between 100-200kHz. Although most of the reduction in emissions seems marginal, mostly in the range of 3-5dB, when converted to linear this is translates into a reduction of more than half.

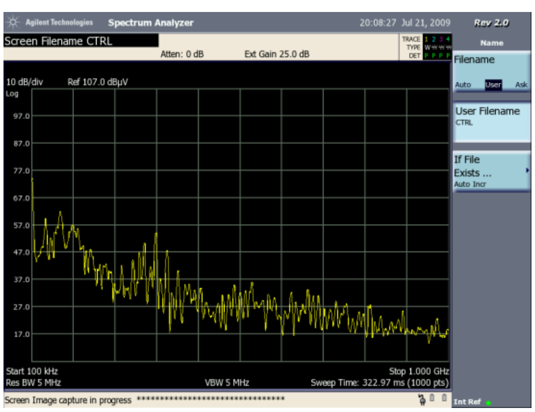


Figure 3.95. Section 1 A&D Ferrites.

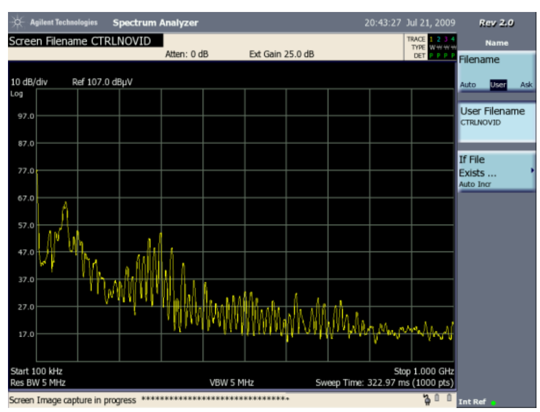


Figure 3.96. Section 1 A&D Ferrites Standby.

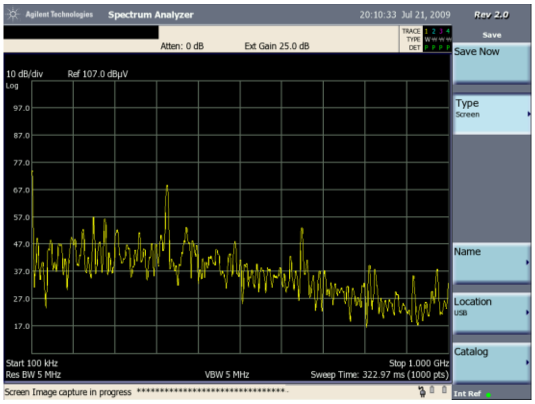


Figure 3.97. Section 2 A&D Ferrites.

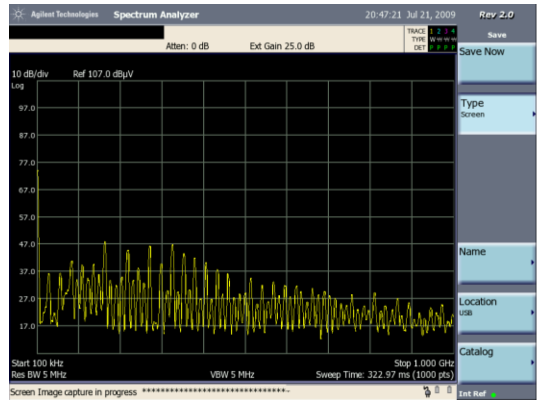


Figure 3.98. Section 2 A&D Ferrites Standby.

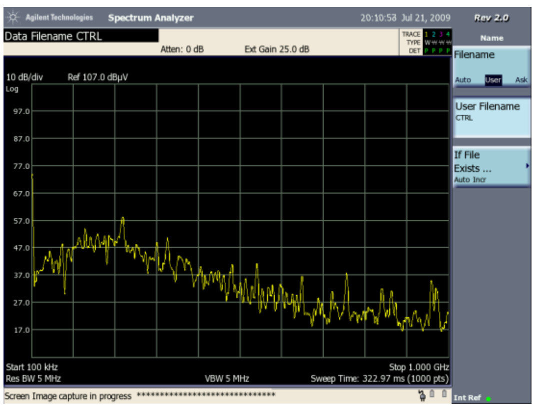


Figure 3.99. Section 3 A&D Ferrites.

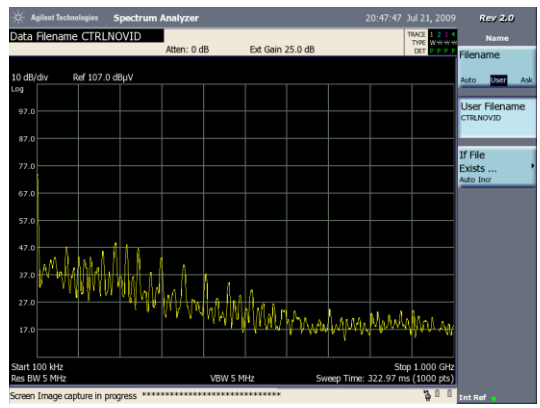


Figure 3.100. Section 3 A&D Ferrites Standby.

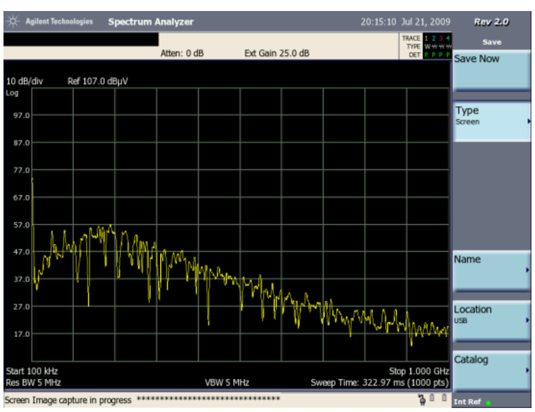


Figure 3.101. Section 4 A&D Ferrites.

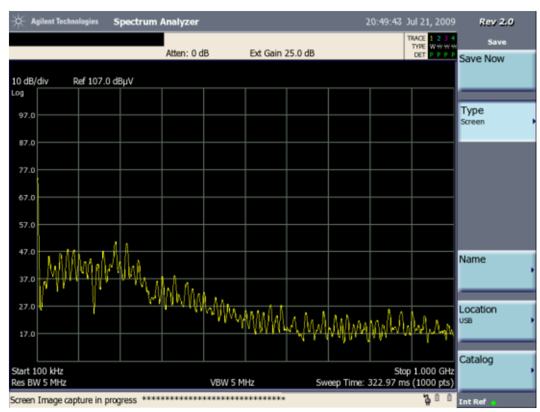


Figure 3.102. Section 4 A&D Ferrites Standby.

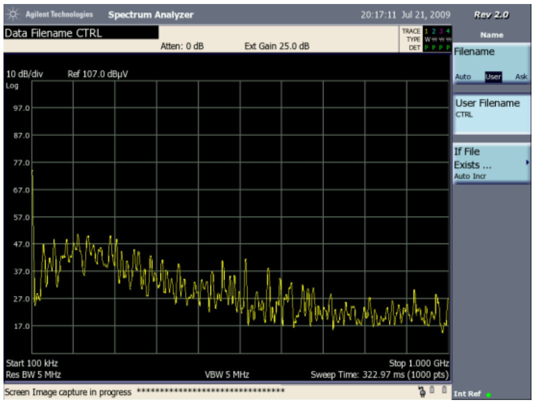


Figure 3.103. Section 5 A&D Ferrites.

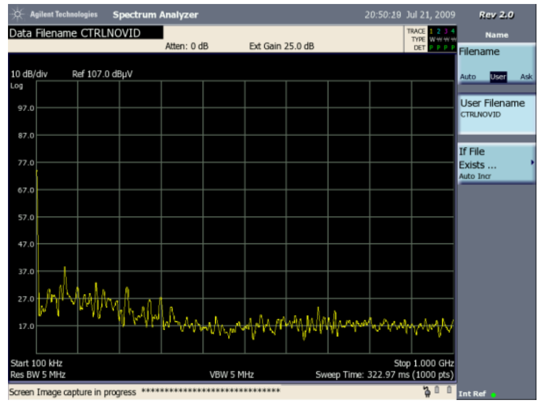


Figure 3.104. Section 5 A&D Ferrites Standby.

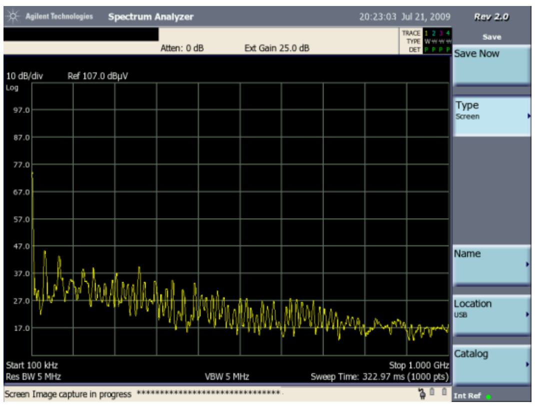


Figure 3.105. Section 6 A&D Ferrites.

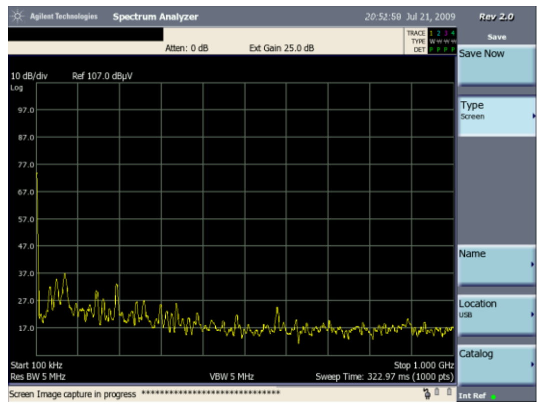


Figure 3.106. Section 6 A&D Ferrites Standby.

The addition of digital ferrites resulted in little reduction in radiated emissions and the result came down to three issues.

1. The analog section radiates significantly more than the digital.
2. The ferrites are not large enough.
3. The board is already optimized.

From Figure 3.95 through Figure 3.106, it can be seen that adding the ferrites to the analog sections did decrease the amount of radiated emissions; however, it cannot be shown that the analog section is radiating significantly more than the digital. This rules out the first issue; next comes the potential that the ferrites are simply too small to be effective. To test this all of the 30Ω ferrites were replaced with 600Ω ferrites on both the analog and digital sections of the PCB. Figures 3.107 through 3.118 show the PCB radiated emissions with 600Ω ferrites installed.

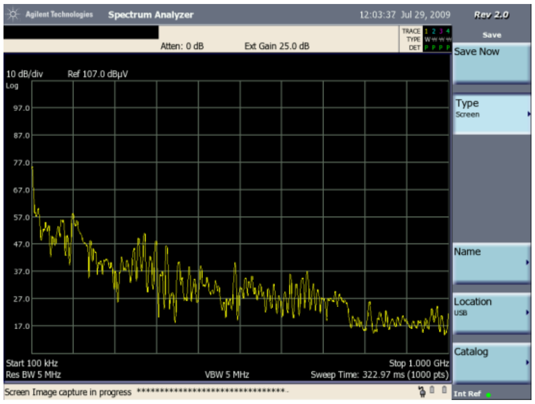


Figure 3.107. Section 1 600Ω Ferrites.

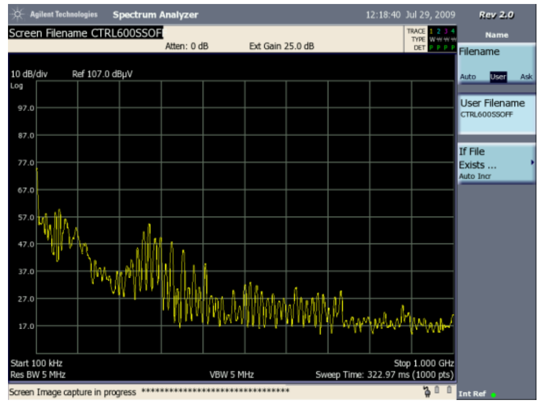


Figure 3.108. Section 1 600Ω Ferrites Standby.

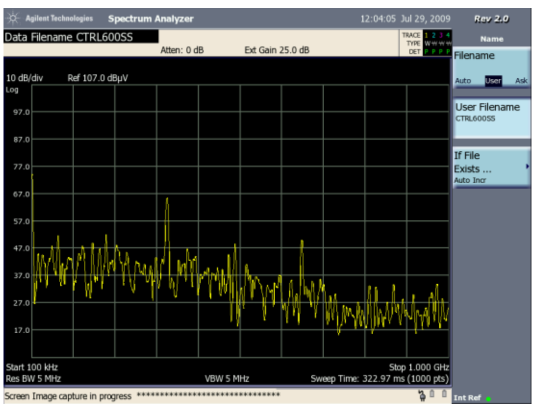


Figure 3.109. Section 2 600Ω Ferrites.

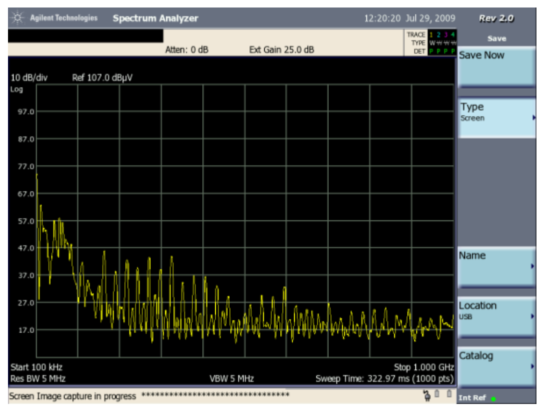


Figure 3.110. Section 2 600Ω Ferrites Standby.

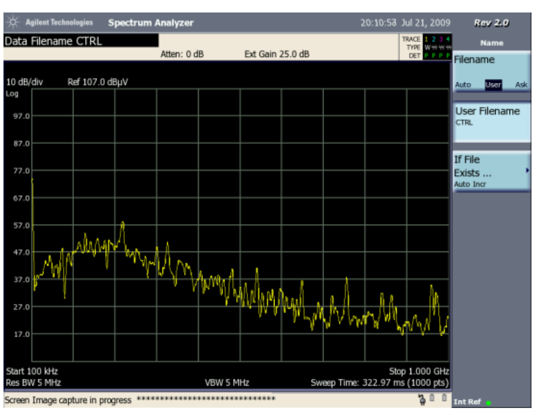


Figure 3.111. Section 3 600Ω Ferrites.

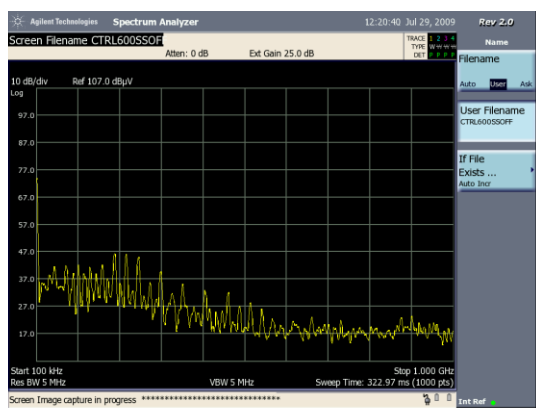


Figure 3.112. Section 3 600Ω Ferrites Standby.

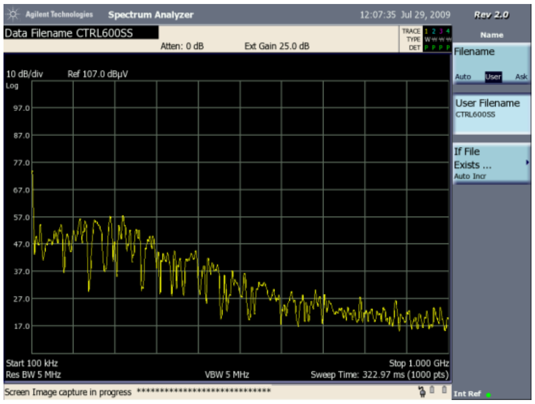


Figure 3.113. Section 4 600Ω Ferrites.

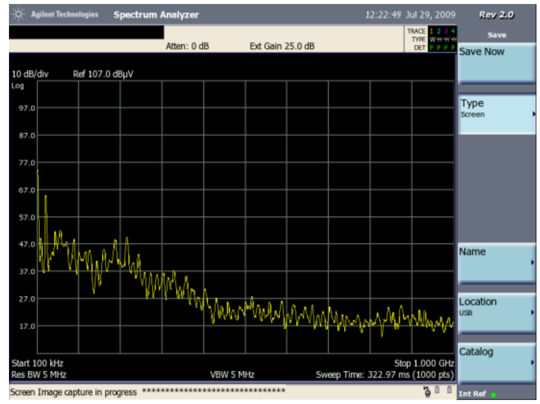


Figure 3.114. Section 4 600Ω Ferrites Standby.

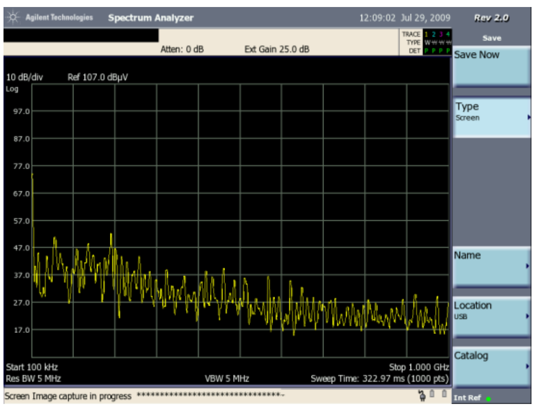


Figure 3.115. Section 5 600Ω Ferrites.

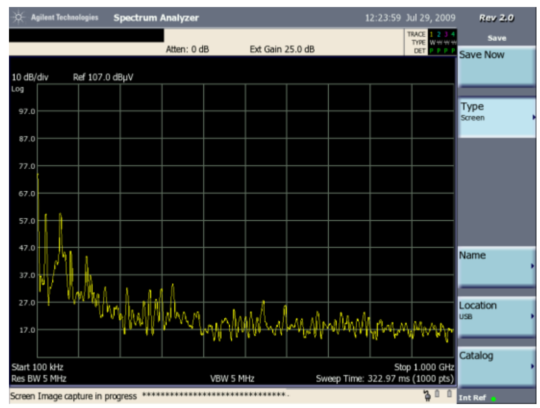


Figure 3.116. Section 5 600Ω Ferrites Standby.

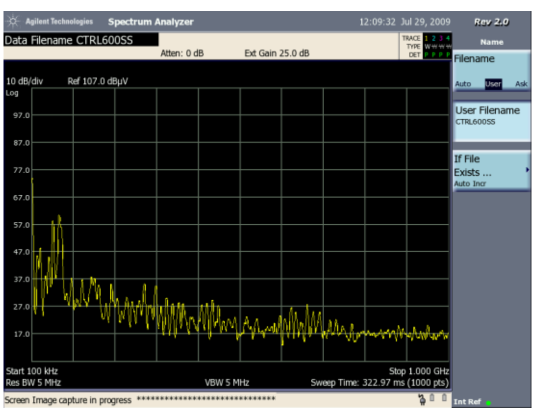


Figure 3.117. Section 6 600Ω Ferrites .

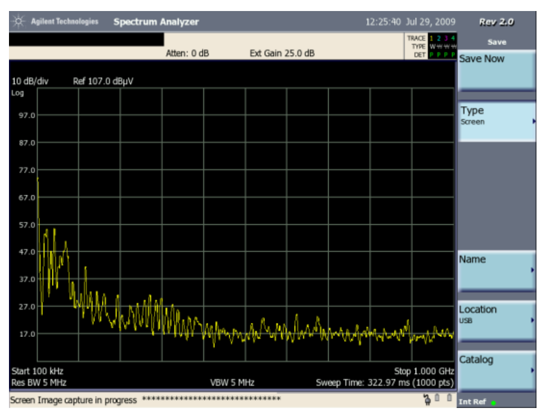


Figure 3.118. Section 6 600Ω Ferrites Standby.

The addition of 600Ω ferrites show a small improvement in radiated emissions. Section 1 (Figure 3.107 and Figure 3.108) show a decrease in radiated emissions over the entire spectrum of several dB when compared to the 30Ω ferrites. However, radiated emissions in Section 5 and 6 (Figures 3.115 through 3.118) radiates significantly more between 100-300kHz indicating there may be a design related issue with the larger ferrites.

To determine if a further reduction in radiated emissions could be achieved by using even larger ferrites, 1500Ω ferrites were installed on the PCB. With the 1500Ω ferrites installed, the operation of the PCB became very intermittent. The size of the ferrite became large enough to prevent some of the high frequency switching components from operating correctly, thus affecting the primary system operation and ruling out the second issue.

It can be determined that the PCB is optimized as best as possible without physically making changes to the layout of the PCB. It was observed that larger ferrites are able to reduce the amount of radiated emissions, and that the largest possible ferrite should be used that doesn't disrupt normal circuit operation. If a ferrite becomes too large it has the potential to make power traces too inductive and disrupt switching digital circuits. This PCB was not tested in a certified FCC testing facility, however if the product were to still fail, other methods could be first attempted prior to redesigning the circuit board. Different shielding methods could be attempted such as placing the PCB in a shielded enclosure or using more expensive shielded cables, both internal and external to the product, in order to pass testing.

CHAPTER 4. SUMMARY AND FUTURE WORK

This study was performed to demonstrate how improperly designed circuits and PCB layouts can have an impact on the amount of radiated EMI from an electronic device. Case studies were presented in which an excessive amount of radiated emissions were able to disturb surrounding electronics. The effects of this disturbance ranged from a mere inconvenience, to a complete failure of potentially lifesaving electronic devices. In order to demonstrate the effects of an improper design and the impact on radiated EMI, individual circuits were designed independently of each other and measured. The improper designs were then evaluated to determine what the cause was for the excessive amount of radiated emissions, and methods presented as possible solutions. Finally, a case study of an electronic device currently on the market was evaluated for radiated emissions. This electronic device was already designed to minimize radiated EMI, but had not been tested in a certified EMI test facility.

In all cases it was found that the improperly designed circuits and PCB layouts radiated more than the control. This indicates that correct signal routing and a good PCB layout is absolutely critical to reduce radiated EMI. From the switching regulator results (Figure 3.37 and 3.38) it was shown that differential or common mode noise due to the switching regulator can propagate throughout the entire system. Depending on how the load is routed, the amount of emissions can increase significantly to where ferrites will have no effect. Providing clean power to all circuits is the most important consideration to reduce both conducted and radiated emissions.

Next, layout techniques to avoid problems such as signal coupling and unintentional antennas must be used, while insuring that impedance matching is implemented in the circuit design and layout. High frequency signals should be manually routed, avoiding the use of an auto-router as much as possible. If an auto-router must be used a thorough check of the finalized PCB should be done. The layout engineer will want to verify that there are no iso-islands on the PCB (copper areas not connected to anything), there are no stub traces, all trace lengths are as short as possible, decoupling capacitors are located as close to the device as possible, the board is properly partitioned into analog/digital/RF sections, and the return path for signal and power routing is directly under the source. Multi-layer PCBs generally perform better than two-layer PCBs due to large ground and power planes that can be placed between layers. Multi-layer PCBs come with a higher price tag though, and with proper routing on a two-layer PCB the same level of emissions can be achieved. High quality cables with shielding will reduce radiated EMI, and prevent EMI from other devices coupling onto it.

The last consideration after the design phase is to evaluate all cabling used within the system. As the frequency of operation increases using the proper cable becomes even more critical because the cable becomes a transmission line, and can turn into an antenna. High quality cables with no ferrites will yield lower radiated emissions than poor quality cables with ferrites on both ends, and proper crimping and construction should always be verified. An improper crimp will allow electromagnetic energy to escape the cable and radiate as EMI.

The prototype case study PCB incorporated and implemented the proper circuit design and PCB layout methods for minimal EMI emissions. The PCB was optimized by using multiple layers and good layout techniques. It was shown that additional ferrites did not have a significant impact. It was found from the case study that a good design will always radiate less without corrections, than a poor design with corrections.

Implementing an EMC expert in the design process to catch possible issues will reduce development costs. The reduction in cost will be due to reducing the amount of design iterations while going through FCC/CISPR testing, and in additional costs due to expensive EMI reducing hardware. Figure 2.16 was presented to show how costs to fix EMI related issues grow exponentially as the product progresses through the development cycle. While the product is in the design phase, as many EMI hardening techniques should be implemented as possible. This often increases the overall price tag of the product, but without passing test the product cannot be sold, and therefore no profit can be made. Once the product has passed test, the hardening methods can be systematically removed from the product to cut cost until the least amount of additional hardware is needed to still maintain compliance.

4.1. Recommended Future Work

Future work for this subject should include additional tests for measuring conducted EMI in an electronic device. An analysis of issues caused by the conducted EMI, and methods to reduce the amount of energy coupling into the device could be performed. Additional measurements could be made using the current radiated EMI examples in a certified EMI testing facility. Correlations between the amount of reduction in radiated emissions measured by the EMI sensing probes, and measured by the testing facility could be found. Due to unforeseen issues, the case study was not able to be tested in a certified testing facility. The design should be tested at a facility while installed in its enclosure, and the EMI troubleshooting flowchart could be used to get the product certified. Finally, additional devices which could potentially cause EMI problems such as spark gaps or motor drivers could be evaluated.

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LIST OF REFERENCES

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APPENDIX

APPENDIX.
ADDITIONAL MEASUREMENT DATA

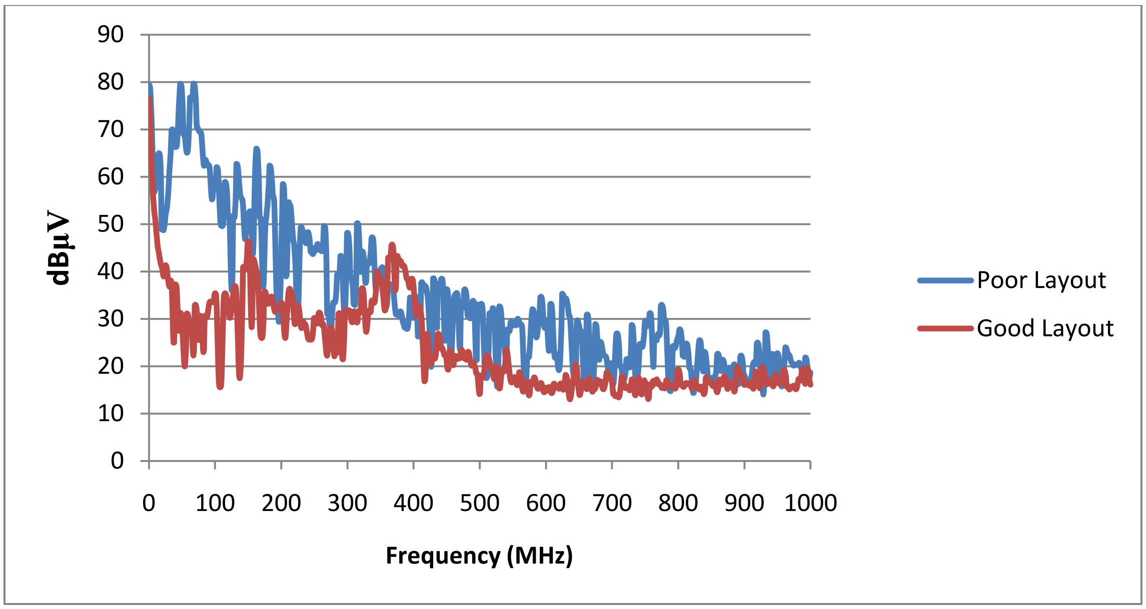

Figure A.1. Common Mode 100Ω Ferrites.

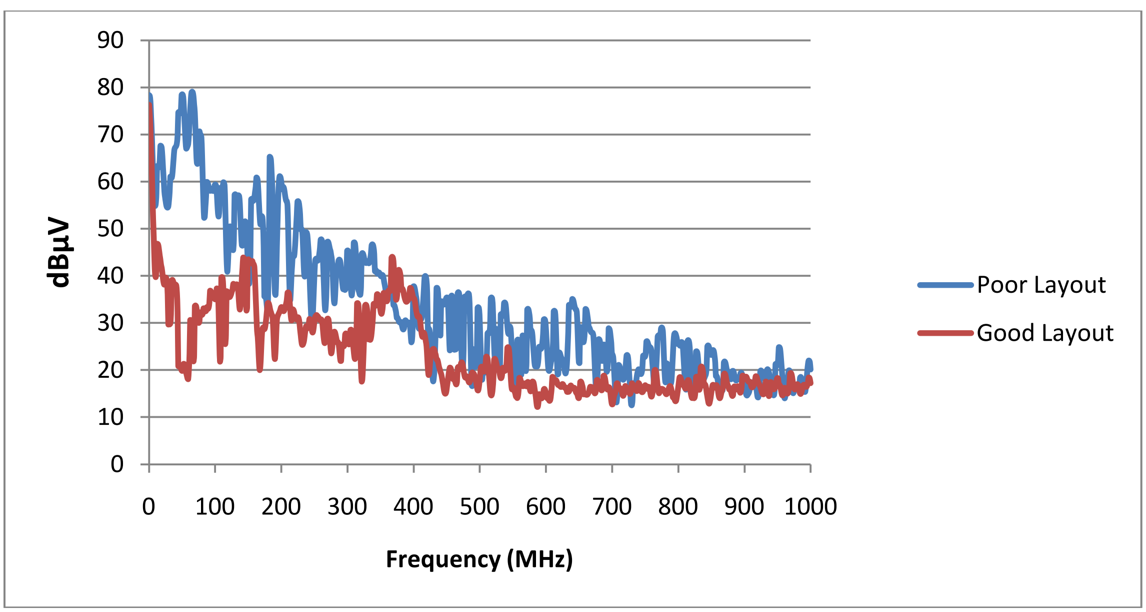

Figure A.2. Common Mode 600Ω Ferrites.

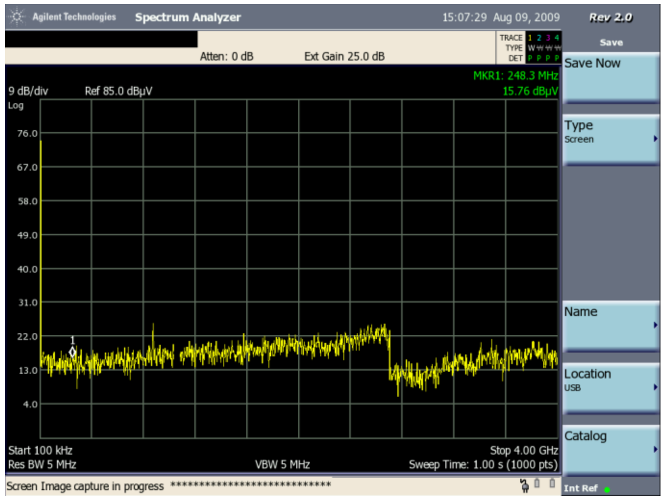

Figure A.3. Good Cable – Radiated at Generator (Input Signal Frequency = 0.25GHz).

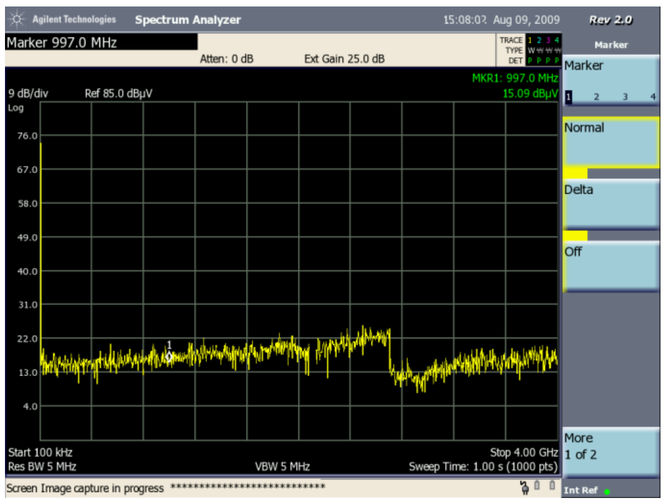

Figure A.4. Good Cable – Radiated at Generator (Input Signal Frequency = 1.0GHz).

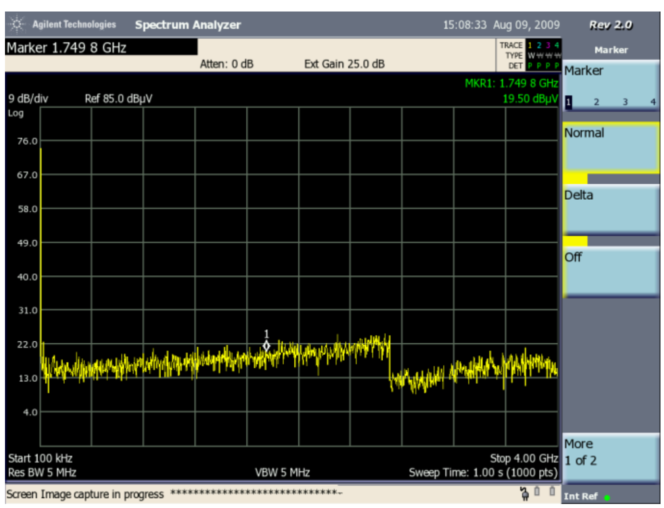

Figure A.5. Good Cable – Radiated at Generator (Input Signal Frequency = 1.75GHz).

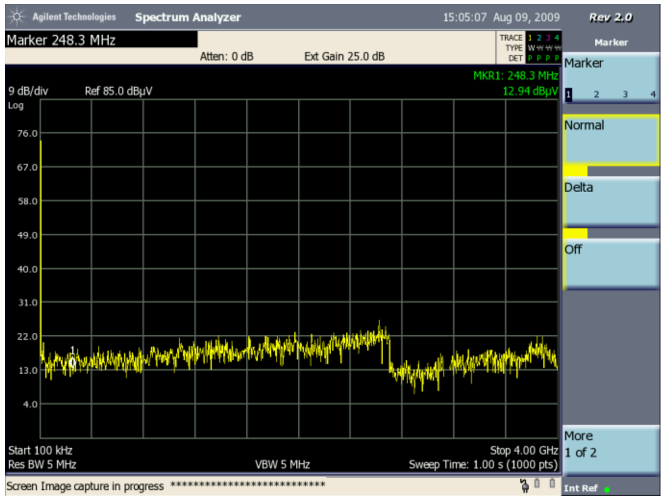

Figure A.6. Good Cable – Radiated at Middle (Input Signal Frequency = 0.25GHz).

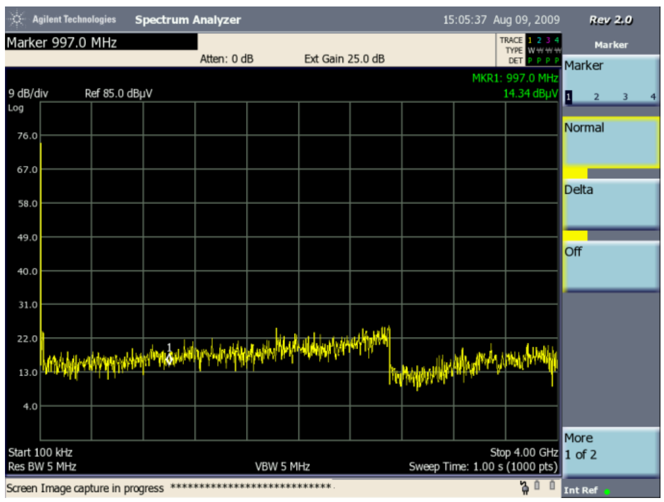

Figure A.7. Good Cable – Radiated at Middle (Input Signal Frequency = 1.0GHz).

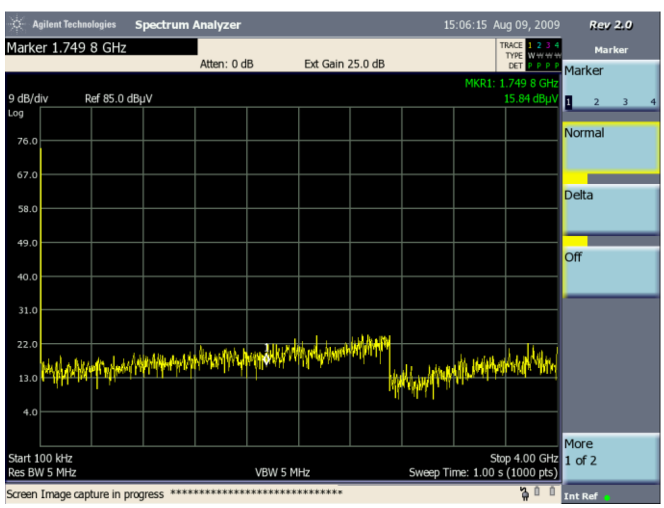

Figure A.8. Good Cable – Radiated at Middle (Input Signal Frequency = 1.75GHz).

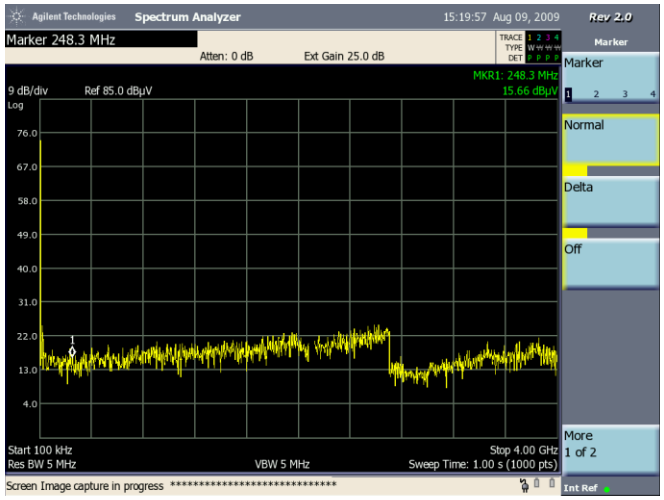

Figure A.9. Damaged Cable – Radiated at Middle (Input Signal Frequency = 0.25GHz).

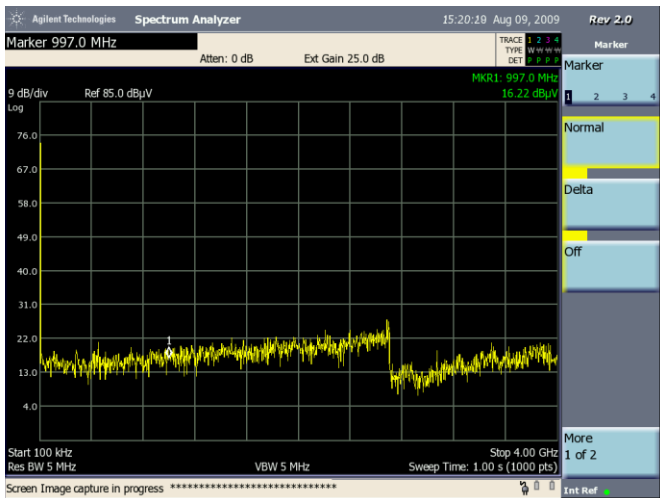

Figure A.10. Damaged Cable – Radiated at Middle (Input Signal Frequency = 1.0GHz).

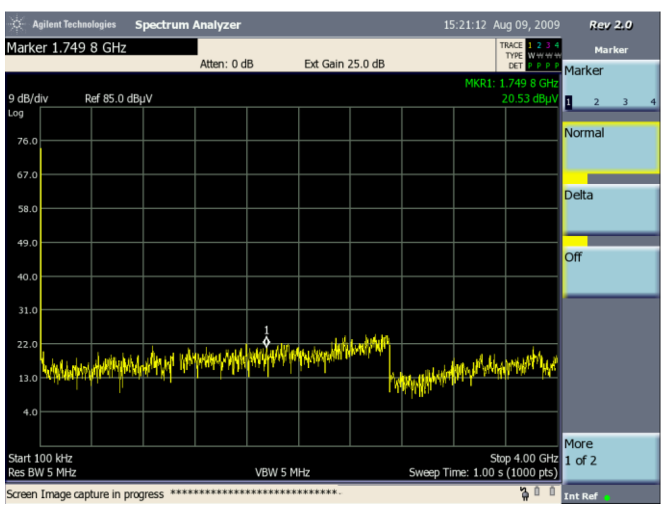

Figure A.11. Damaged Cable – Radiated at Middle (Input Signal Frequency = 1.75GHz).

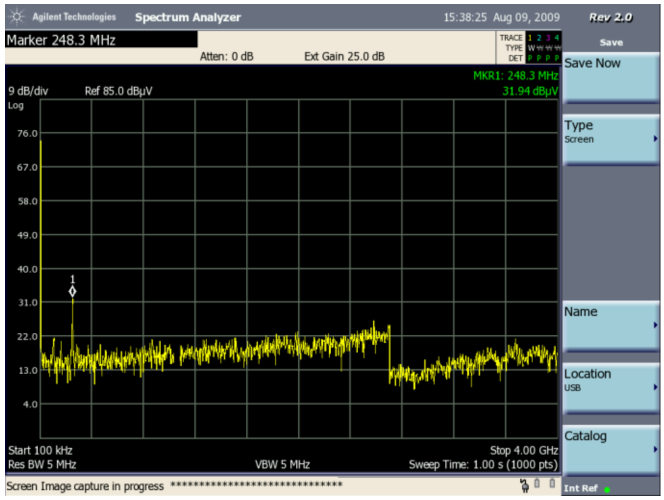

Figure A.12. Broken Cable – Radiated at Middle (Input Signal Frequency = 0.25GHz).

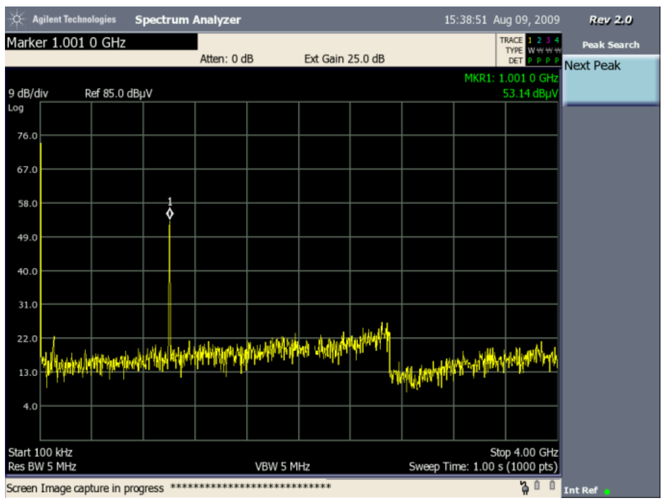

Figure A.13. Broken Cable – Radiated at Middle (Input Signal Frequency = 1.0GHz).

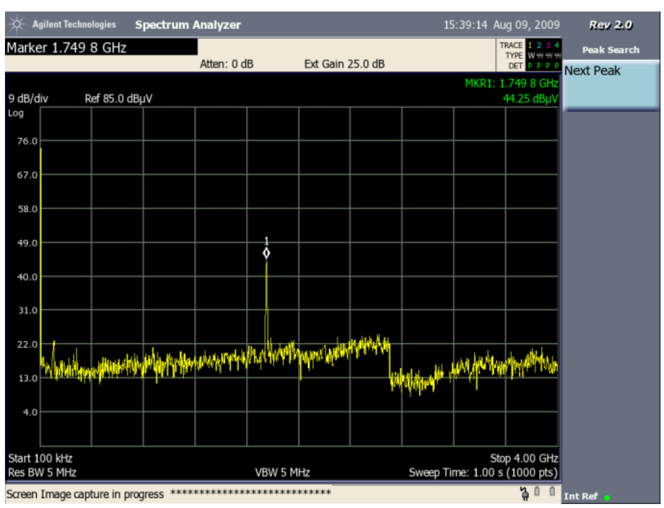

Figure A.14. Broken Cable – Radiated at Middle (Input Signal Frequency = 1.75GHz).

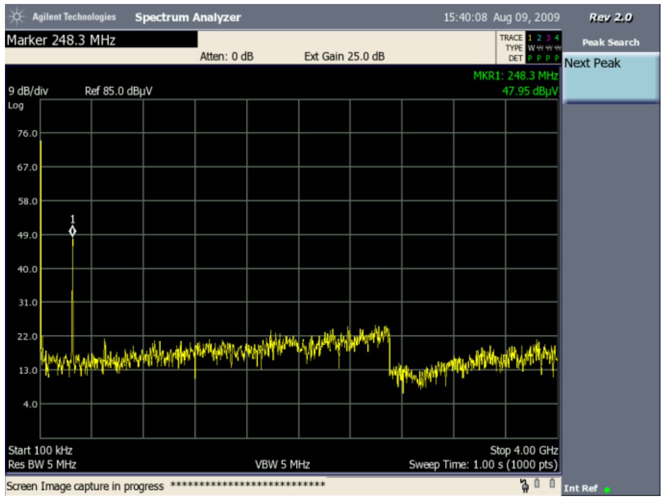

Figure A.15. Broken Cable – Radiated at Generator (Input Signal Frequency = 0.25GHz).

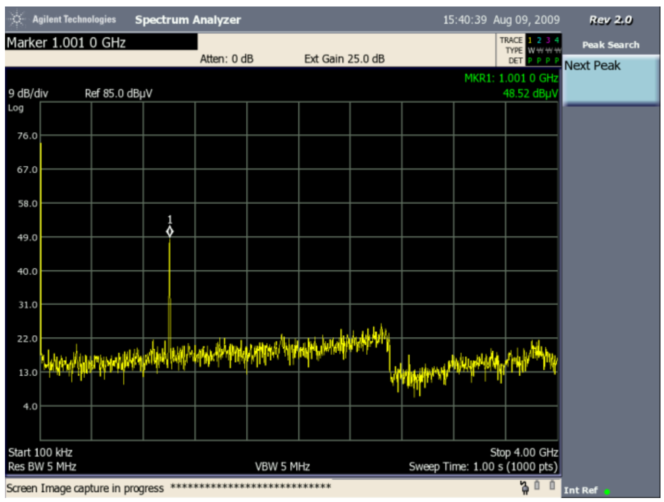

Figure A.16. Broken Cable – Radiated at Generator (Input Signal Frequency = 1.0GHz).

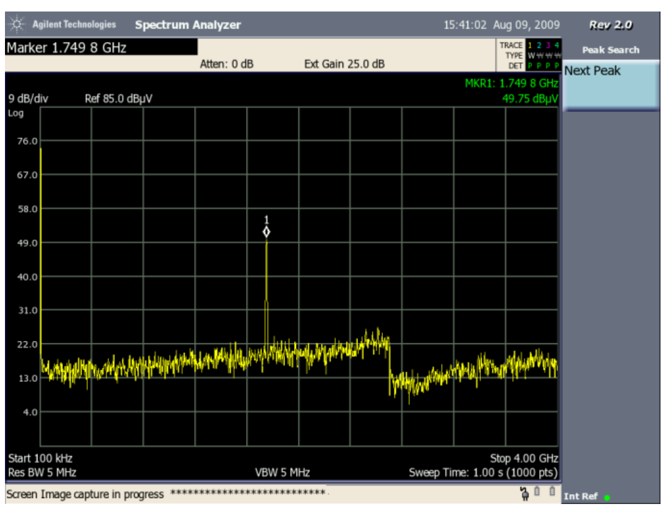

Figure A.17. Broken Cable – Radiated at Generator (Input Signal Frequency = 1.75GHz).