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A Selective History of Part 15 of the FCC Rules

An Engineering Perspective

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FEATURE ARTICLES

SPECIAL COVERAGE

COLUMNS

DEPARTMENTS

FDA's New AI Tool Struggles with Basic Tasks

The U.S. Food and Drug Administration (FDA) is already encountering implementation issues with its efforts to leverage the potential benefits of AI-based technologies.

NBC News reports that the FDA's newly launched AI tool, named Elsa, is already having challenges addressing even simple tasks, such as uploading documents or answering questions from users. Further, the report says that the agency's AI tool is not yet connected to the Internet, limiting its ability to access newly published studies. NBC's sources say that Elsa "still needs significant work" to become operational.

The FDA initially announced that it would launch Elsa to help agency scientific reviewers and investigators "work more efficiently" by accelerating clinical protocol review and shortening the timeline for the execution of scientific evaluation.

UK Overhauls Post-Market Surveillance Requirements for Medical Devices

The United Kingdom has recently implemented strict regulations applicable to the post-market surveillance (PMS) of medical devices.

According to a press release on the country's government website, the new PMS regulations will require device manufacturers to proactively track the performance and safety of their products. The new regulations, which took effect on June 16th, apply to all UKCA- and CE-marked devices placed on the market after June 16th, including in vitro diagnostic devices (IVDs), active implantable medical devices, and other medical device technologies intended for use in healthcare settings and in-home care.

Specifically, the new PMS regulations will require medical device manufacturers to:

- Gather and assess data on how their medical devices perform in daily use, to help increase their ability to detect safety and performance issues;
- Expand incident reporting, including serious incidents relating to side effects associated with the use of their medical device;
- Leverage new data analysis reporting options to support earlier detection of trends; and
- More proactively assess and manage risks and promptly notify users when safety issues arise.

FCC Streamlines Cable Rate Regulations

The U.S. Federal Communications Commission (FCC) has taken the next step in its effort to streamline its regulations by approving new rules applicable to cable rates for television services.

In a Report and Order, the Commission implemented a number of changes originally proposed in 2018, including the removal of 77 rules or requirements and eight forms the FCC deems "unnecessary." Specific changes include:

- Deregulating cable equipment that is not used exclusively to receive the basic level of cable service;
- Deregulating small cable systems serving 15,000 or fewer subscribers;
- Declining to extend rate regulation to commercial establishments; and
- Revising and simplifying the process for cable operators to establish an initial regulated rate for services.

The latest action by the FCC is part of "Delete, Delete, Delete," the Commission's deregulation initiative, which it says is intended to "facilitate network modernization, infrastructure development, and performance innovation."

Updated EU Regulations Introduce Repairability Scores

To help promote the anticipated life of smartphones, tablets, and other consumer technology devices, the European Union (EU) Commission has introduced a new way for consumers to assess the extent to which their devices are repairable.

The EU's new "repairability score system" will soon be displayed on the EU's new Energy Label that accompanies electronic devices. According to the Commission, "the repairability scores provide a clear and easy-to-understand rating of a product's repairability from A (highest) to E (lowest)." Individual repairability scores consider several critical factors, including key product components, the steps and tools needed to disable a device and access replaceable components, and the availability of replacement information and spare parts.

The Commission says that the new repairability score system will provide consumers with information essential to understanding the environmental sustainability of new devices when they purchase them, allowing them to make more informed choices.

Penn State Students Rewrite Kirchhoff's Law

Nothing lasts forever... Even something 165 years old!

A student research team at Pennsylvania State University has reportedly demonstrated a "dramatic breaking" of the long-standing Kirchhoff's law of thermal radiation.

According to an article posted on the Penn State website, the team and their academic advisors designed a structure utilizing nonreciprocal emitters that can send emissions in different directions. Their breakthrough directly counters the long-standing premise that materials that absorb electromagnetic radiation at a given wavelength and angle must be equal to their capability to emit at the same wavelength and angle.

Zhenong Zhang, a doctoral candidate in mechanical engineering at Penn State, says that the team's findings could be used to make the harvesting of solar energy even more efficient. "If we have nonreciprocal emitters," says Zhang, "we can send emissions toward a different direction. Then, we could place another solar cell there to absorb this part of energy, increasing the overall power conversion efficiency."

Linxiao Zhu, an assistant professor of mechanical engineering at Penn State, expanded further on the team's research. "We designed a structure that has five semiconductor layers, each with slightly different compositions…the structure absorbs and emits thermal radiation over multiple wavelengths, so we expect to see the effect over a broad wavelength band."

The results of the work conducted by the team were published in *Physics* Magazine.

A SELECTIVE HISTORY OF PART 15 OF THE FCC RULES: AN ENGINEERING PERSPECTIVE

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By Daniel D. Hoolihan

From the time that Heinrich Rudolf Hertz first demonstrated the transfer of electrical energy from one antenna to another in the late 1880s, humanity has witnessed, observed, and enjoyed the fruits of electromagnetic radiated fields. The electromagnetic waves first produced by Hertz in his lab in 1886 were proof that James Clerk Maxwell's 1864 theory of "electromagnetic waves" was correct.

Hertz published a series of papers in the last years of the 1880s that verified the characteristics of the "Hertzian Waves" with respect to frequency, amplitude, speed (velocity of light), and other physical parameters. Hertz never realized the practical importance of his discovery and did not explore the applications of the "Hertzian Waves," which became known as "radio waves" over time.

However, many premier scientists of the day did recognize the importance of the discovery, including Guglielmo Marconi, Nikola Tesla, Edwin Armstrong, Lee DeForest, and many others.

Wireless telegraphy was the first commercial use of the "radio waves" proven to exist by Hertz. The Boer War in the late 1890s showcased the first military use of wireless telegraphy; both warring nations used it to communicate between the units of their armed forces in that war.

Experiments continued in the 1890s and early 1900s to extend the range of the "radio waves" until the first messages were successfully sent from Europe to the United States.

As radio technology developed, radios became more popular and more prevalent, leading to interference between radio transmitters. To control the interference, countries met to develop an agreement to control the interference since the radio waves did not recognize or respect country geographical boundaries. The result was a "Final Protocol" signed at the Berlin Conference in 1903 that stated that "services be organized in such a way as to avoid interference with other stations." This agreement was further strengthened in a second agreement signed at the Berlin conference in 1906.

The first radio law passed by the United States was the Wireless Ship Act of 1910, which required ocean-going vessels to be equipped with "an efficient apparatus for radio communication."

TITANIC DISASTER - 1912

Marconi successfully put radio stations on ships so that they could communicate with shore-based stations and with one another. Marconi's company even installed radios on the RMS Titanic, which sank during its maiden voyage after hitting an iceberg on the night of April 14-15, 1912. However, despite the availability of this advanced technology, the Titanic could not communicate with the Californian, a ship only ten miles away. As a result, many lives were lost that could have been saved. (The radio operator on the Californian had gone to bed and shut his radio system off about one hour before the Titanic hit an iceberg).

The Carpathia, one of the ships that heard the Titanic's SOS call and came to save the survivors of the Titanic, had to come from fifty-eight miles away, and it took several hours for it to arrive. Anyone thrown or left in the cold water after the Titanic sank did not survive long enough to be rescued. People in the Titanic's lifeboats or left clinging to wooden pieces of furniture/fixtures were rescued by the sailors of the Carpathia.

As a result of the Titanic disaster, the U.S. Congress passed the Radio Act of 1912, which dictated that each ocean-going ship had to have a radio room that

In December 1921, only two wavelengths were set aside in the U.S. for radio stations intending to broadcast to a general audience. The number of broadcasting stations in the U.S. grew rapidly and, by the end of 1922, more than 500 stations were operating in the country.

was in operation 24 hours a day. This same legislation put the Department of Commerce in charge of licensing radio transmitters in the U.S., which consisted primarily of maritime (ship radio stations) and amateur radio stations.

COMMERCIAL RADIO - 1920s

In December 1921, only two wavelengths were set aside in the U.S. for radio stations intending to broadcast to a general audience. The first wavelength was 360 meters (833 kHz), and it was designated for "entertainment," while the second wavelength was 485 meters (619 kHz) and could be used for "market and weather reports." The number of broadcasting stations in the U.S. grew rapidly and, by the end of 1922, more than 500 stations were operating in the country. The number of reserved transmitting frequencies also grew until they filled the frequencies from 550 kHz to 1500 kHz in ten-kHz steps.

Radio Act of 1927

Interference problems and other administrative challenges that arose with the rapid expansion of commercial (broadcast) radio forced Congress to enact additional legislation that was published as the Radio Act of 1927. This act created the Federal Radio Commission. Its five-member panel was given the power to grant and deny licenses, assign station frequencies and power levels, and issue fines for violations associated with the 732 broadcasting stations then in existence.

Communications Act of 1934

To handle the increasing judicial and technical issues surrounding commercial amplitude modulated (AM) radio, Congress passed the Communications Act of 1934. This act abolished the Federal Radio Commission and transferred its power to the Federal Communications Commission (FCC). We still operate today under this 1934 Act which has been amended and added to many times.

PART 15 FIRST APPEARS - 1938

In 1938, we see the first mention of low-power devices in Part 15 of the FCC Rules. FCC Chief Engineer, Ewell Jett, argued that if certain radio frequency emissions were sufficiently weak and short-ranged so as not to be considered measurable, they would not rise to the level of harmful interference. One example of a typical device meeting these requirements was a Philco Radio and Television Corporation miniature transmitter remote-control device for its standard broadcast receivers. It was the company's belief that the transmitter could be operated without a radio station license since it operated by using an individual's own home radio to receive and retransmit the low-level signal.

After studying the device, the FCC proposed that the operation of low-power transmitters without a license would be allowed if they met proposed standards that were so low in amplitude that they could not interfere with interstate communications. The new rules were called "the Low Power Rules." These new rules required the use of minimum power and precautions against interference, plus the radiated field was limited to 15 microvolts per meter (uV/m) at a distance equal to the wavelength divided by 2 pi. These rules were adopted in November 1938 and remained in effect for low-power communication devices until 1957.

REVISION OF PART 15 - 1949-1954

After World War II, low-power radio devices increased rapidly, and some of the newer devices were designed to use frequencies that were higher than the AM broadcast band. To address these new issues, the FCC instituted a proceeding in April 1949 to address an overall revision of Part 15.

It wasn't until April 1954 that the FCC issued a Notice of Proposed Rulemaking (Docket No. 9288) in which low-power devices were divided into two categories: 1) incidental radiation devices in which

The 1970s will be known for the release of a proposal to limit electromagnetic emissions from computers. Computers were classified as incidental radiation devices, and they were covered by Part 15 requirements.

the RF energy is generated as a by-product of normal operation; and 2) restricted radiation devices in which RF energy is generated deliberately. Based on replies from industry, the FCC released new rules in December 1955, which established a table of permissible radiation limits for all receivers operating (tuning) between 30 and 890 megahertz.

A second report and order in Docket No. 9288 was issued in 1956 and 1957. It addressed community antenna television receivers and low-power communication devices (miniature transmitters).

PART 15 IN THE 1960s AND 1970s

With respect to Part 15, the 1970s will be known for the release of a proposal to limit electromagnetic emissions from computers. Computers were classified as incidental radiation devices, and they were covered by Part 15 requirements. But the computer manufacturers were not designing or manufacturing their computers to any emission criteria in the late 1960s and early 1970s.

Led by the Computer and Business Equipment Manufacturers Association (CBEMA), the computer industry started meeting with the FCC to discuss the technical situation. The leading computer manufacturer at the time was IBM, and the other major U. S. manufacturers were Control Data Corporation, Univac, Digital Equipment Corporation, Honeywell, National Cash Register (NCR), Burroughs, and Xerox. The meetings were usually attended by one EMC engineer from each company, except for IBM, which sent two engineers and a lawyer.

The FCC started the discussions due to an increasing number of complaints regarding interference to radio and TV signals from computers. As the number of home computers (personal computers) began to increase in the late 1970s, the meetings between the computer industry and the FCC became more frequent.

FCC Docket No. 20780, which was released in 1976 to address restricted radiation devices, was the vehicle the FCC used to develop the case for limits on emissions from computers.

The FCC rules for computers defined two different classes of equipment: Class A for commercial computers and Class B for personal/home computers. The limits for Class A equipment were three times more relaxed than those applicable to home computers.

The FCC said that…

> *"It was initiating this rulemaking proceeding because it had found that the general requirement for restricted radiation devices – section 15.7 – adopted in 1938 and carried over into Part 15 in 1948, were no longer suitable for the numerous devices now regulated by these rules. Therefore, it proposed to clarify and further define these requirements, to recognize and define certain specific devices and in a few cases, to establish new requirements."*

Based on the studies done and released by CBEMA, the FCC adopted new rules on digital electronic equipment (especially computers) in September 1979.

COMPUTER LIMITATIONS ON INTERFERENCE - 1980s

An order adopted by the FCC in March 1980 made a few minor changes to the original rules on computers and moved the effective dates of the rules to October 1981 for new equipment and October 1983 for older equipment. Under the revised rules, manufacturers had the responsibility to design and build their computers to limit their electromagnetic emissions to satisfy the regulatory limits set by the FCC. In addition, warning statements had to be placed on the computer and in the accompanying instruction manuals concerning their interference potential.

The FCC rules for computers defined two different classes of equipment: Class A for commercial computers and Class B for personal/home computers. The limits for Class A equipment were three times more relaxed than those applicable to home computers (in technical terms, about 10 dB more relaxed), resulting in higher limits.

TELECOM CERTIFICATION BODIES (TCBS) - 1990s - 2000s

Another major impact of FCC Part 15 devices on society started in the last years of the 20th century. FCC General Docket Report and Order 98-68 was released at the end of 1998. This Docket was further defined in FCC Publication Notice DA-00-1223, which was issued in August 1999. This Notice signaled the implementation of what were called Telecom Certification Bodies (TCBs), beginning in June 2000.

FCC Public Notice DA 99-1640 was released in 1999. It stated:

> *"FCC Provides Further Information on the Accreditation Requirements for Telecommunication Certification Bodies (TCBs). This Public Notice provided further information on the accreditation requirements for TCBs including information on the accreditation body, the National Institute of Standards and Technology (NIST) of the Department of Commerce. The Notice concluded that "TCBs will have a major impact on how equipment manufacturers get their products approved in the future. The telecom industry representatives are looking forward to faster turnaround times on their product approvals."*

TCBs are private industry-independent organizations that have been authorized under the FCC program to issue grants to electronic product manufacturers for the certification of specific types of telecommunication products covered under the program's scope. This included Part 15 products such as computers and computer peripherals. The program began with 13 designated TCBs.

The importance of the TCB program was further enhanced by the release of ET Docket No. 03-201 (FCC 04-1665). One of the FCC Commissioners at that time had the following to say about the Docket:

> **Separate Statement of FCC Commissioner Jonathan S. Adelstein**
> *Re: Modification of Parts 2 and 15 of the Commission's Rules for Unlicensed Devices and Equipment Approval; ET Docket No. 03-201*
>
> *"The development of wireless ISPs and the advent of so-called hotspots using unlicensed spectrum has been one of the Commission's great success stories over the last*

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Model Number	Frequency Range	Power dBm	Gain dB
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AMP10140	10 MHz-6.0 GHz	33	33
LNA70001	50 MHz-6.0 GHz	25	25
MPA10135	0.2-20.0 GHz	33	33
MPA10134	0.5-26.5 GHz	30	30
MPA10136	1.0-18.0 GHz	30	30
MPA10132	2.0-8.0 GHz	33	33
LNA70008	2.0-18.0 GHz	20	20
MPA10133	4.0-8.0 GHz	33	40
LNA70013	18.0-40.0 GHz	10	20
AMP30040	18.0-26.5 GHz	34	36
AMP30039	26.5-40.0 GHz	30	36
AMP40048	18.0-40.0 GHz	27	30
AMP40002	41.0-47.0 GHz	33	33

Popular Models

Model Number	Frequency Range	Power dBm	Gain dB
AMP20093	10 kHz-250 MHz	300	55
AMP20095	10 kHz-250 MHz	600	58
AMP20081	80-1000 MHz	500	57
AMP20083	80-1000 MHz	750	60
AMP20079	0.7-6.0 GHz	100	50
AMP20080	1.0-6.0 GHz	200	53
AMP20098	2.0-8.0 GHz	120	51
AMP20102	2.0-8.0 GHz	400	56
AMP20072	6.0-18.0 GHz	300	55
AMP20154	6.0-18.0 GHz	700	58
AMP40013	18.0-26.5 GHz	10	40
AMP40028	18.0-26.5 GHz	150	52
AMP40038	26.5-40.0 GHz	20	43
AMP40035	26.5-40.0 GHz	120	52
AMP20005	18.0-40.0 GHz	25	44
AMP20167	18.0-40.0 GHz	60	48
AMP40049	40.0-50.0 GHz	1	30
AMP20057	1.0-2.0 GHz	4 KW	66
AMP20144	2.0-4.0 GHz	4 KW	66
AMP20097	4.0-8.0 GHz	2 KW	63
AMP40053	8.0-12.0 GHz	2 KW	63

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several years. I support this item because it continues our efforts to promote the development of unlicensed devices and services. The tremendous growth of WiFi in the 2.4 GHz band was facilitated by the licensing (or more appropriately the "unlicensing") approach initially adopted by the Commission for this band. Part 15 of our rules allows manufacturers to develop technologies for the unlicensed-bands that anyone can use without a license. We must continue this policy approach so that we encourage as many avenues or technologies as possible for broadband and other important services to reach consumers, no matter where they live.

"My goal as a policymaker is to maximize the services and information that flow over our airwaves. A regulatory framework for innovation can provide the necessary conditions that support the growth and development of spectrum-based services, including continued use of the unlicensed bands. Such a framework functions in a manner akin to a greenhouse, in which plants are protected from the elements by a structure and are nurtured so that they can thrive on their own within it.

"I believe that in the NPRM adopted today, the Commission properly strives for such an approach to spectrum management. We want to enhance our existing Part 15 structure so that it continues to encourage the growth of the unlicensed industries, but also controls the elements, like minimizing interference that may impact existing and future operators. Just as a greenhouse can support different types of plant forms, our framework for innovation does the same – it must be flexible enough to accommodate all different kinds of technologies, such as those used with the latest antennas. Our framework does not choose which technology will survive, and which will not, but it must create an environment that allows the different seeds of technology to truly have an opportunity to grow and develop on their own. I believe that we have such an approach here, and I am optimistic that our framework for innovation will enable new technologies in the unlicensed space to continue to meet the public's demand for broadband more efficiently."

By 2015, the FCC had turned over all Certification testing to the TCBs.

Part 15 – Low Power Devices continues to be an important part of the FCC Rules. Many clever innovations have resulted from the category of "unlicensed" products unleashed on society by smart and hard-working electronic engineers.

OTHER 1999 FCC ACTIONS

FCC Public Notice DA 99-890 was released in May 1999. The FCC's Office of Engineering and Technology (OET) clarified certain procedures that allow manufacturers to market central processing unit (CPU) boards as separate components.

In its FCC ET Docket No. 99-231 – Notice of Proposed Rulemaking, the FCC amended Part 15 of its rules relative to spread spectrum devices. It proposed changing the rules for frequency hopping devices operating in the 2.4 GHz band with a maximum output power of 1 Watt to allow for wider operational bandwidths.

Wireless Devices

The rapid expansion of wireless devices (initially driven by the release of the Apple iPhone in 2007) proved that the FCC was forecasting the future demand of testing appropriately. The FCC, as a single source of testing, would not have been able to test all the new product designs released as a result of the wireless demand. However, industry responded with additional testing labs and TCBs to handle the avalanche of wireless products produced by the innovative electronics industry.

SUMMARY

Part 15 – Low Power Devices continues to be an important part of the FCC Rules. Many clever innovations have resulted from the category of "unlicensed" products unleashed on society by smart and hard-working electronics engineers. The FCC has responded by opening new swaths of spectrum at higher and higher frequencies to allow the development of additional devices with expanded capabilities. The FCC has worked closely with TCBs to ensure the effectiveness of the Part 15 Rules, including informal rulings called KDBs (for Knowledge Database) guidelines for qualifying products to the FCC Part 15 Rules.

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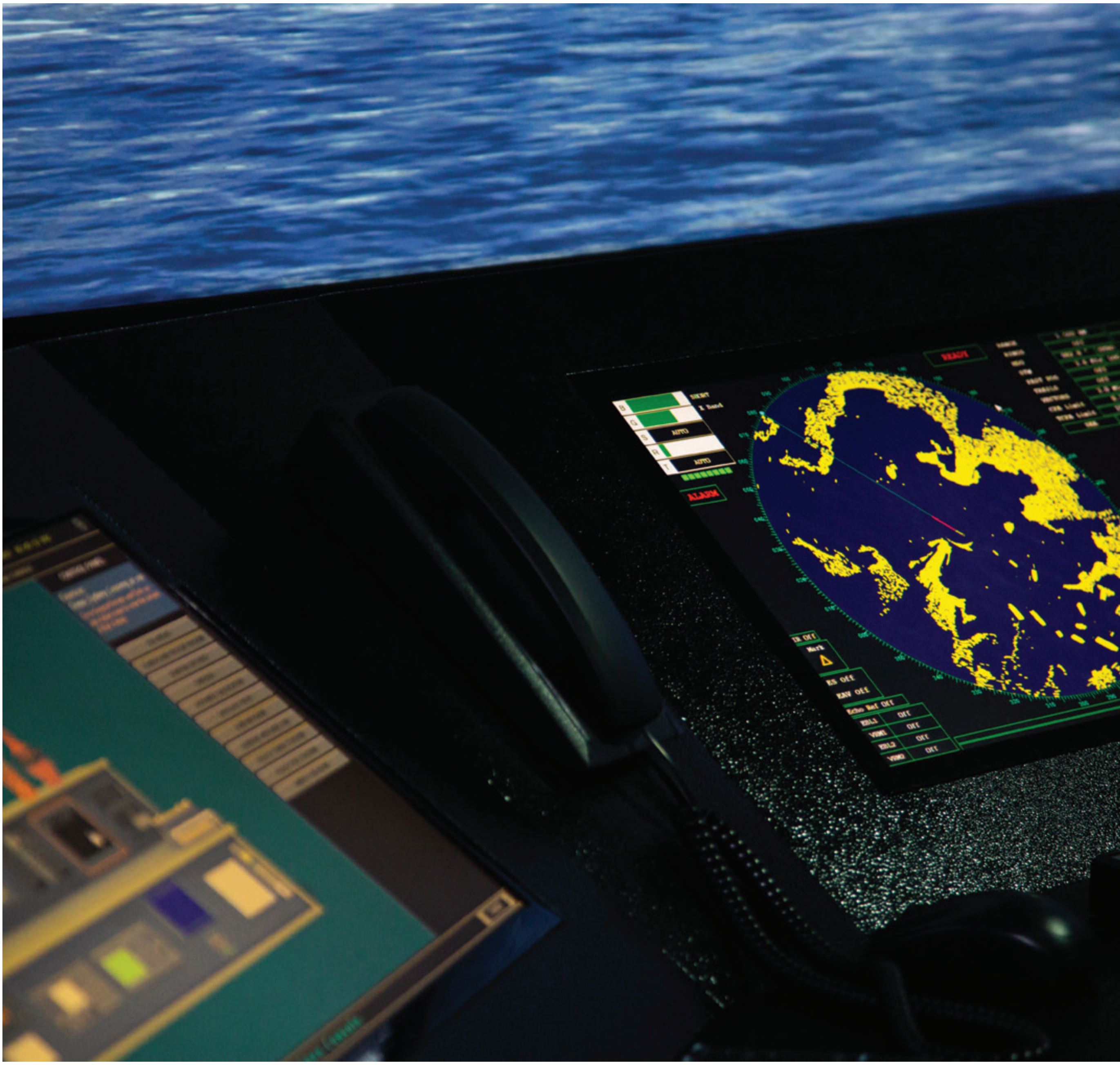
STATISTICS OF ELECTROMAGNETIC FIELDS WITHIN WIRE-COUPLED, NESTED REVERBERANT ENCLOSURES



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By Marshall D. Sowell, Kyle A. Shea, and Carl E. Hager IV

Editor's Note: *The paper on which this article is based was originally presented at the 2024 IEEE International Symposium on Electromagnetic Compatibility & Signal/Power Integrity (EMC + SIPI), where it received recognition as the Best Symposium Paper.*

Naval platforms present the most severe Electromagnetic Environment (EME) in the world, created by high-powered electromagnetic emitters such as air-search, surface-search, fire-control, and navigation radars, as well as broadband communications and electronic warfare systems. The limited space available aboard surface platforms necessitates that these high-powered electromagnetic emitters be located in proximity to other electronic/electrical systems aboard ships. Sensitive electronic equipment is typically contained within Radio Frequency (RF) reflective cavities (i.e., below-deck spaces, equipment enclosures, etc.) for protection from harsh EMEs. Communications and power cables are typically routed between enclosures within such spaces, introducing inadvertent coupling paths between cavities. Additionally, as the topside EME continues to increase and as below-deck transmitters (i.e., RFID, Wi-Fi, etc.) are incrementally installed within the Fleet, understanding and predicting the field distributions within coupled spaces will assist in characterizing electronic equipment performance and assuring personnel safety.

Statistical electromagnetic formalisms for electromagnetic fields within RF-reflective cavities had their beginnings through the study and use of reverberation chambers (RCs) for Electromagnetic Compatibility (EMC) testing [1] and are now widely accepted within the EMC community as a tool for compatibility and susceptibility testing [2-4]. RCs are electromagnetically reflective cavities with a high quality (Q) factor where the fields excited within the cavity *reverberate* [4-6]. The addition of tuners within the chamber (also known as paddles or stirrers) allows for the electromagnetic (EM) boundary conditions to be easily changed so the EM fields can be perturbed discretely (mode-tuned) or continuously (mode-stirred).

The statistics of RCs have been well studied and show the magnitude of a single spatial electric-field component, $|Er|$, will follow a chi-distribution with two degrees of freedom, and the square magnitude of this same component, $|Er|^2$, follows a chi-square distribution with two degrees of freedom; both distributions are also widely known as the Rayleigh and exponential distribution, respectively [4, 6].

This project takes an alternative approach to the already growing interest in describing the field statistics within nested cavities [8-13]. Previous studies examined the case when two cavities are coupled via an *aperture* [8-9]. These studies hypothesized that the double-Rayleigh distribution would model the nested field statistics, but were unsuccessful in demonstrating it as a good fit for all test cases. This study hypothesized the same distribution but alternatively uses N wire penetrations as the coupling mechanism (vice apertures) and analyzes the statistics of the received power for each N case.

STATISTICAL BACKGROUND

Hypothesized Distributions

The chi-squared distribution with two degrees of freedom, also known as the exponential distribution and the double-Rayleigh distribution, will be considered in this study ([8] provides further background for both distributions). Because the *double-Rayleigh* distribution is the product of two Rayleigh random variables (RVs), which implies studying electric fields, the authors have chosen to name its power form [8] the *double-exponential* (DE) distribution because it implies the studying of received

Two RCs were used in this study, one constructed of aluminum and the other constructed of zinc-coated steel. The aluminum cavity is referred to as the nested enclosure, while the zinc-coated steel cavity is referred to as the external enclosure.

powers. (This distribution is not to be confused with the Laplace distribution that is sometimes referred to as a "double-exponential distribution" [14-15].) The probability density function (PDF) for the exponential distribution and double-exponential distribution is

$$f_X(x) = \frac{1}{\mu} e^{-\frac{x}{\mu}}, \qquad x \geq 0 \qquad \text{Eq. 1}$$

and

$$f_Z(z) = \frac{2}{\xi} K_0\left(2\sqrt{\frac{z}{\xi}}\right), \qquad z \geq 0 \qquad \text{Eq. 2}$$

respectively, where μ is the scale parameter in (1), $x = |Er|^2$, ξ is the scale parameter in (2), Z is the product of two independent exponential RVs ($Z = XY$), and KO(~) is the 0^{th} order modified Bessel function of the second kind. See [8] for further explanation and the cumulative distribution function (CDF) of (2).

Scale Parameter Estimation and Hypothesis Testing

The scale parameter for equations (1) and (2) was estimated using the maximum likelihood estimation (MLE) method. Anderson-Darling (AD) goodness-of-fit (GOF) testing was used to assess the data against an exponential distribution and chi-squared GOF testing was used to assess the data against both an exponential distribution and the DE distribution. Detailed mechanics of MLE when applied to the exponential distribution and the DE distribution, as well as why these two GOF tests were selected are discussed in detail in [8].

EXPERIMENTAL SETUP

Reverberant Cavities

Two RCs were used in this study, one constructed of aluminum and the other constructed of zinc-coated steel. The aluminum cavity is referred to as the nested enclosure, while the zinc-coated steel cavity is referred to as the external enclosure. The physical dimensions of both enclosures are provided in Table 1. The external enclosure contains both a vertical and horizontal Z-fold tuner, while the nested enclosure contains a vertical Z-fold tuner. The nested enclosure was placed on support foam within the working volume of the external enclosure. Figures 1 and 2 provide a sketch and picture of the nested measurement setup for a single coupling antenna (N = 1). The single coupling antenna is circled in red in Figure 2.

Dimensions	Nested Cavity (meters)	External Cavity (meters)
Length	0.70	5.51
Width	0.44	5.25
Height	0.31	3.56

Table 1: Enclosure dimensions

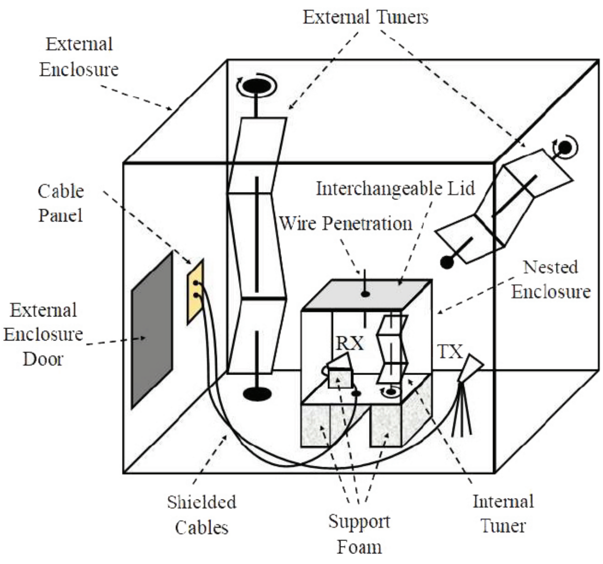

Figure 1: Nested cavity schematic with N = 1 wire penetration

Figure 2: Nested cavity setup with N = 1 wire penetration

Measurement Setup

A Vector Network Analyzer (VNA) was used for data capture in this study. The VNA output power was set to 0 dBm, and a 150W amplifier was used to ensure all coupled measurement configurations achieved data above the VNA's noise floor. S21 measurements were swept across the frequency range of 8 to 12 GHz at 1 MHz steps (4001 points total) with a step dwell sweep setting of 1 millisecond (ms). This frequency range ensured both enclosures were sufficiently overmoded. All three tuners (two external, one nested) were simultaneously stepped 100 times. At each step, an |S21| measurement was performed across the full frequency range, resulting in 4001 sample sets (one at each frequency), each set containing 100 independent samples.

Both enclosures were originally measured in isolation from one another to verify that both indicate exponential received power. The AD test was used to test each sample set against an exponential distribution at a 95% confidence level. The null hypothesis rejected 4.75% of the sample sets for the external enclosure and 6.15% of the sample sets for the nested enclosure, both at the defined confidence level. The rejection rates are near that of an ideally operating RC at a 95% confidence and can therefore be used for this study.

Test Cases

A visual representation of the nested configuration is shown in Figure 1 and a picture of the setup is shown in Figure 2. As depicted in Figure 3 on page 20, a single insulated 12-gauge wire, measuring six inches in length, evenly extends externally and internally to the nested enclosure through a 3.5-millimeter (mm) hole. The wire does not make any electrical contact with the enclosure. This single

Figure 3: Nested enclosure setup with N = 1 wire penetration

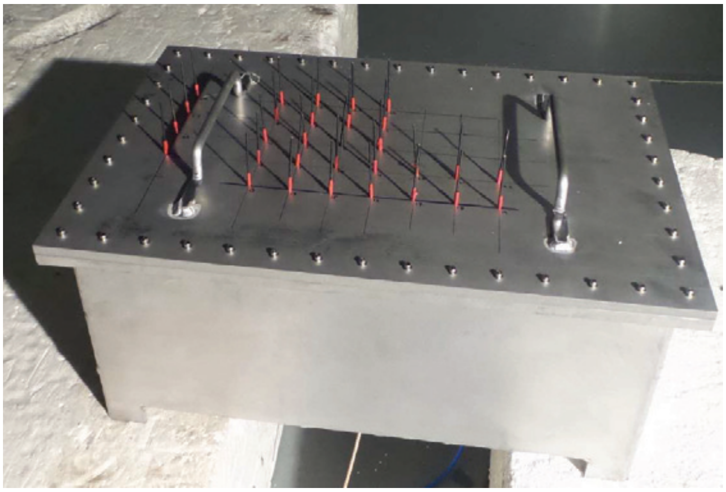

Figure 4: Nested enclosure with N = 30 wire penetrations

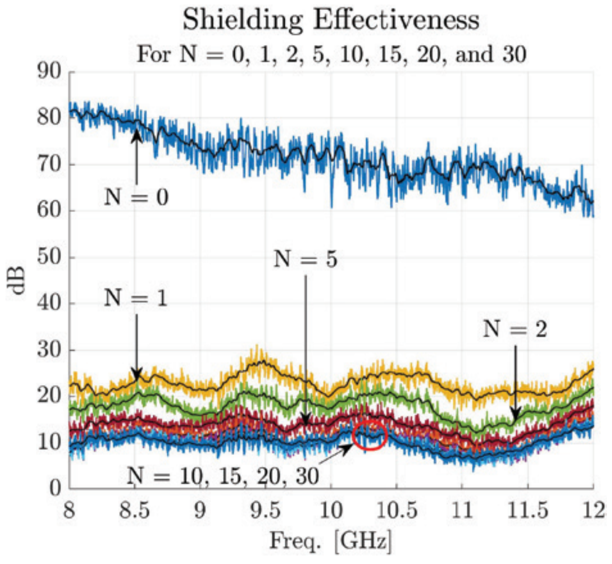

Figure 5: SE for different test cases

wire penetration is the N = 1 test case. A total of thirty test cases were performed, each N case corresponding to the number of independent wire penetrations. The N = 1, 2, 5, 10, 15, 20, and 30 test cases are shown in this paper for brevity; Figure 4 shows the N = 30 wire penetrations, each separated by ~5-centimeters. For each test case, the total number of independent distributions tested was 4001, one for each frequency recorded. The ideal GOF test would have a 5% rejection rate at a 95% confidence (which is approximately 200 distributions rejecting the null hypothesis).

Shielding Effectiveness

The nested enclosure's shielding effectiveness (SE) was measured following the procedures in [16] and is shown in Figure 5. The nested enclosure was measured to show approximately 80 dB of SE at the lower frequencies, which reduces to 60 dB SE at the higher frequencies over the frequency range of this study (N = 0 configuration). The other seven test cases are also shown in Figure 5, and it shows the SE significantly reduces for the N = 1 wire penetration, and as more wire penetrations are added the impact of the reduction in SE reduces eventually reaching a point of diminishing returns, i.e., the two enclosures were acting as a single enclosure. The black line for each test case represents a 50MHz smoothing to emphasize the trend the SE follows for each test case.

DE GOF RESULTS

Table 2 shows the total percentage of values that rejected the null hypothesis in GOF testing, where each column represents the associated distribution and respective GOF test. All GOF testing was used to test the data against the respective hypothesized distribution at a 95% confidence. Data tested against an exponential distribution using the

AD test and the chi-square GOF test are represented by EX-A^D and EX-χ^2, and data tested against the DE distribution using the chi-square GOF test is represented by the DE-χ^2.

For the N = 1 case, the exponential distribution is almost entirely rejected for both the AD and chi-square GOF tests, with the total amount rejecting the null hypothesis being 99.90% and 97.30%, respectively. However, the total amount of distributions that rejected the DE distribution was 5.52%, strongly indicating that the DE provides a good fit for the data in this specific test case.

For the N = 2 and 5 cases, the exponential distribution is rejected decreasingly less than the N = 1 case, but still at too great a rate for evidence of an exponential distribution. However, comparing the N = 2 and 5 cases, it is becoming evident that fewer distributions

Test Case N	Percentage of Rejected Distributions Tested at a 95% Confidence Value		
	EX-A^D	*EX-χ^2*	*DE-χ^2*
1	99.90 %	97.30 %	5.52 %
2	89.85 %	56.46 %	24.29 %
5	39.04 %	14.55 %	63.06 %
10	13.72 %	6.57 %	79.76 %
15	9.92 %	6.92 %	84.68 %
20	7.50 %	6.65 %	86.50 %
30	6.52 %	6.35 %	88.25 %

Table 2: Percentage of exponential and DE distributions that rejected the null hypothesis at a 95% confidence

can reject the null hypothesis when tested against the exponential distribution. Additionally, it should not be surprising that the AD and chi-square GOF rejections are different, with the chi-square being less powerful than the AD test, as shown in [23]. The rejection rate for the DE distribution demonstrates the exact opposite trend as the exponential distribution, with more distributions being rejected as the number of independent wire penetrations increases.

As N continues to increase, the evidence of the exponential distribution increases to where finally the two GOF tests have near identical total rejected values (i.e., N = 30 test case). For the N ≥ 20 test cases, the two enclosures are statistically operating as a single RC. Alternatively, the DE distribution only provides a good fit for a single wire penetration (i.e., N = 1) and is rejected for all other test cases (i.e., N > 1).

The transitions in GOF rejection rates are visually depicted in Figure 6. As N increases for the exponential distribution, the total number of rejected values decreases, as seen in Figure 6a for the AD GOF tests, while the total number of rejected values increases for the DE distribution using the chi-square GOF test shown in Figure 6b.

N-COUPLED CAVITY HYPOTHESIS

We have shown that for a single wire penetration between nested enclosures, the received power closely follows the DE distribution. However, as N coupling mechanisms increase, the DE is no longer suitable to describe the distribution of the fields, and the exponential distribution only begins to confidently describe the fields for sufficiently high N cases. We now examine the potential for a new hypothesized distribution to be able to describe this *transitional region* between the two extremes (i.e., 1 < N < 30).

Taking a step back and conceptually understanding what is going on *under the lid*, each new coupling mechanism (i.e., wire penetration) is introducing an independent exponential RV into the nested enclosure. Since each wire penetration is introducing power into the nested enclosure through the coupled power from the exposed wire within the external enclosure, each exponential RV is from the same parent distribution, specifically the external chamber's exponential distribution. Additionally, when sufficiently spaced (by at least λ/4 to ensure spatially uncorrelated samples), each wire penetration introduces an independent, identically distributed (IID) RV.

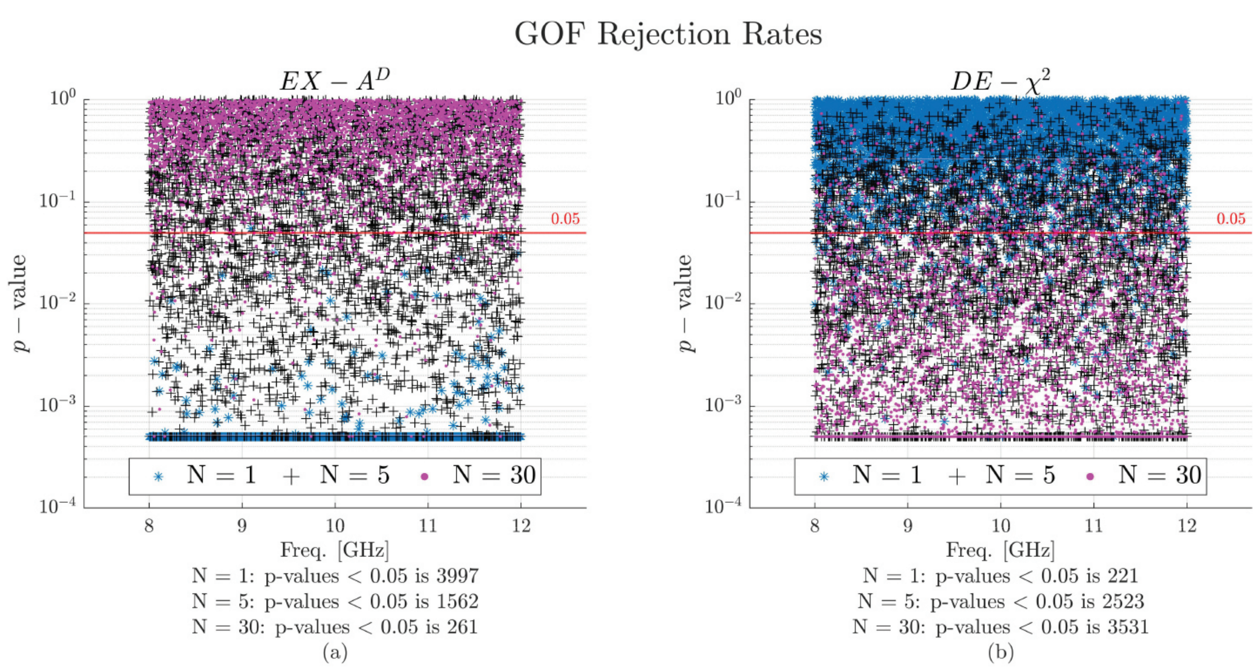

Figure 6a and 6b: GOF rejection rates for the N = 1,5, and 30 test cases. Data tested at a 95% confidence. Data tested against distribution (Figure 6a) and a DE distribution (Figure 6b).

The Erlang Distribution

Based on the characteristics of the physical mechanics of the electromagnetic fields within the nested enclosure, one may predict that the distribution of the total power coupled into the nested enclosure from the external enclosure may be sufficiently described by the sum of N IID exponential distributions, or equivalently the Erlang distribution [17-18]. The Erlang distribution is a special case of the gamma distribution, specifically for positive integer values of N [18-19], and the PDF of the Erlang distribution is defined as

$$f_W(w) = \frac{w^{N-1} e^{-\frac{w}{\sigma}}}{\sigma^N (N-1)!}, \qquad w, \sigma \geq 0 \qquad \text{Eq. 3}$$

where $W = \Sigma_N X_N$ (the sum of N IID exponential RVs), N is an integer, and σ is the scale parameter (σ = 1/λ in [18], where λ is the rate parameter).

Nested Enclosure Hypothesized Distribution for N ≥ 1

With the total power introduced into the nested enclosure defined by the Erlang distribution, following the same principles defined in [8], the distribution of the nested cavity is defined as the product of the Erlang RV (W) and the nested enclosure's exponential RV (Y)

$$Z' = WY \qquad \text{Eq. 4}$$

where Z' is the RV of a distribution the authors have termed the *modified*-Double Exponential (MDE) distribution. By invoking a statistical theorem for the product of two independent RVs [20],

$$f_{Z'}(z') = \int_{-\infty}^{\infty} f_W(w;\sigma) f_Y\left(\frac{z'}{w};\mu\right) \left|\frac{1}{w}\right| dw \qquad \text{Eq. 5}$$

For all N test cases, the chi-square GOF test rejects the null hypothesis at a near-ideal rejection rate. This indicates that the MDE distribution provides a good fit when describing the statistics of received power within the nested enclosure for all quantities of wire penetrations tested.

where $f(w; \sigma)$ and *(inline equation)* represent the Erlang and exponential distributions, respectively. We derive the PDF of the MDE through leveraging 3.471.9 in [21], and changing the lower bound in (5) to 0 because all variables are positive values.

$$f_{Z'}(z') = \frac{2}{\xi(N-1)!}\left(\frac{z'}{\xi}\right)^{\frac{N-1}{2}} K_{N-1}\left(2\sqrt{\frac{z'}{\xi}}\right), \quad z' \geq 0$$

Eq. 6

where ξ is the scale parameter and $\xi > 0$. The MDE CDF is then calculated through the integration of the PDF, or specifically,

$$F_{Z'}(z') = 1 - \frac{2}{(N-1)!}\left(\frac{z'}{\xi}\right)^{\frac{N}{2}} K_N\left(2\sqrt{\frac{z'}{\xi}}\right)$$

Eq. 7

In the specific case of one independent wire penetration (i.e., N = 1), it is easily shown that the PDF and CDF of the MDE reduce to the DE PDF in (2) and CDF in [8]. By employing MLE to estimate the MDE parameter ξ, the log likelihood function is derived to be (8), and the same numerical approach described in [8] was used to estimate the scale parameter ξ.

MDE GOF TEST RESULTS

The chi-square GOF test is again used to test the data against the hypothesized MDE distribution at a 95% confidence for all N test cases. Table 3 repeats the data in Table 2 with a new column, identified by MDE-χ^2, reflecting the percentage of distributions that rejected the null hypothesis at the 95% confidence for the hypothesized MDE distribution using the chi-square GOF test. For all N test cases, the chi-square GOF test rejects the null hypothesis at a near-ideal rejection rate.

Test Case N	Percentage of Rejected Distributions Tested at a 95% Confidence Value			
	EX-A^D	EX-χ^2	DE-χ^2	MDE-χ^2
1	99.90%	97.30%	5.52%	5.52%
2	89.85%	56.46%	24.29%	5.45%
5	39.04%	14.55%	63.06%	5.97%
10	13.72%	6.57%	79.76%	5.50%
15	9.92%	6.92%	84.68%	6.12%
20	7.50%	6.65%	86.50%	6.17%
30	6.52%	6.35%	88.25%	5.62%

Table 3: Percentage of MDE distributions that rejected the null hypothesis at a 95% confidence level

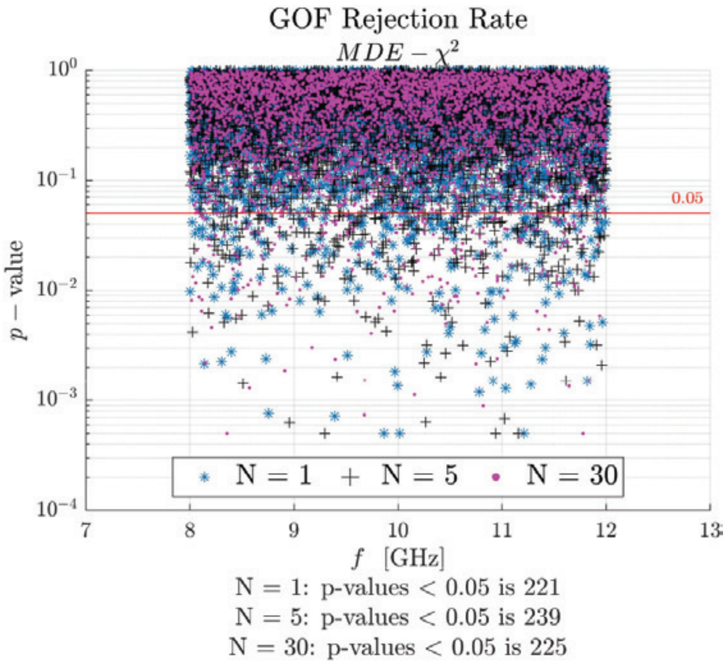

Figure 7: GOF rejection rate for data tested against the MDE distribution. Data tested at a 95% confidence.

This indicates that the MDE distribution provides a good fit when describing the statistics of received power within the nested enclosure for all quantities of wire penetrations tested. Figure 7 shows the p-values for the N = 1, 5, and 30 test cases and the total number of p-values below 0.05, further emphasizing that the MDE distribution is a good fit to the experimental data.

For the N = 30 test case, it is interesting that both the exponential and MDE provide a good fit to the experimental data. The authors intuitively interpret this result from the following: for large N, the Erlang distribution converges to a Normal distribution by the Central Limit Theorem [20]. (In [11], the authors reach a similar conclusion for "the case of many apertures.") Additionally, by the Law of Large Numbers, the variance of the Normal distribution approaches zero with increasing N [18, 20], or equivalently approaches a Dirac delta function (a constant) [24]. Figure 8 illustrates both the mean normalized Erlang distribution converging to a Normal distribution and the variance of the Normal distribution decreasing with increasing N for theoretical mean normalized data. Therefore, for large N, (4) approaches the product of a "constant" and the exponentially distributed random variable of the nested enclosure, or simply a scaled exponential distribution.

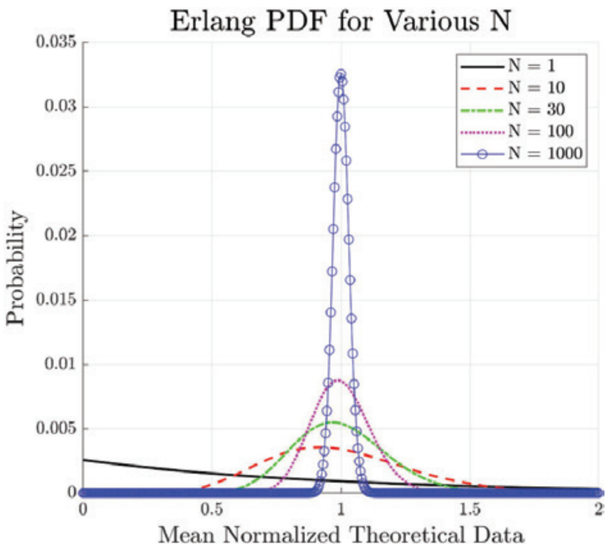

Figure 8: Theoretical mean normalized Erlang PDF for various N with the Erlang scale parameter set to 1.

The authors recognize that the derivation of the MDE (and DE) does not account for energy returning to the external enclosure through the wire penetration(s). Hence, the MDE may not sufficiently describe the statistics for all enclosure/coupling configurations. For example, at lower frequencies where the wall losses are less dominant than antennas [6] more energy may be exchanged between the external and nested enclosures through various coupling mechanisms. Therefore, the authors are working to derive a distribution by applying a conservation of energy approach similar to [22], which was done for averages.

$$\ln\left(L(z_i', \xi)\right) = \sum_{i=1}^{n} \left\{ \ln(2) - \ln(\xi) - \ln\left((N-1)!\right) + \frac{N-1}{2}\left(\ln(z') - \ln(\xi)\right) + \ln\left(K_{N-1}\left(2\sqrt{\frac{z'}{\xi}} \right) \right) \right\} \quad \text{Eq. 8}$$

CONCLUSION

A study has been performed in which two analyses have been completed to characterize the statistical distribution of a nested enclosure when coupled by N wire penetrations. Mode tuning was utilized to simultaneously perturb the electromagnetic environments in both external and nested enclosures. Both enclosures were first shown to exhibit exponentially distributed received power when tuned in isolation. Appropriate and rigorous statistical tests were used to evaluate the GOF for three hypothesized distributions: the exponential, DE, and MDE distributions.

The DE distribution provided a good fit when the two enclosures were coupled by a single wire penetration ($N = 1$). For the $N > 1$ cases, non-DE behavior was observed. The characteristics of an exponential distribution became evident as a large number of wire penetrations ($N \geq 20$) were implemented on the enclosure. For $N < 20$ wire penetrations, the disagreement between the rejection rates of the AD and chi-square GOF tests against the exponential was too different and too large to suggest that exponential behavior is present at the 95% confidence level.

The MDE distribution was then introduced and shown to have a good fit for all test cases ($1 \leq N \leq 30$), indicating it is a good distribution to choose when characterizing the received power under the conditions present in this experiment. Further work will be done to analyze nested enclosures when energy balance [22] is taken into consideration, as well as when apertures are used as test cases.

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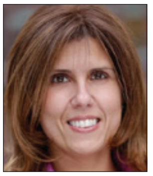
Booth 429

SECURING THE PATIENTS: CYBERSECURITY STANDARDS FOR CONNECTED MEDICAL DEVICES

Mitigating the Risk of Cyber Attacks in Remote Patient Care with IEEE SA Cybersecurity Standards

Maria Palombini is the Healthcare & Life Sciences Global Practice Lead at the IEEE Standards Association (IEEE SA). Palombini can be reached at m.palombini@ieee.org



By Maria Palombini

In the age of digital health, connected medical devices are transforming patient care. From insulin pumps and glucose monitors to smart inhalers and wearable ECGs, these devices deliver real-time data, enable remote monitoring, and improve clinical outcomes. However, as connectivity increases across devices, so does the risk of cyber attacks. For engineers tasked with designing these devices, security by design is more imperative than ever for compliance.

THE EXPANDING THREAT LANDSCAPE

Cybersecurity threats targeting connected medical devices are growing in frequency and sophistication.[1] These devices often operate in complex environments, rely on wireless communication, and store sensitive patient data. A single vulnerability can compromise both device functionality and patient safety and privacy.

Consider a scenario in which a hacker exploits a vulnerability in a Bluetooth-enabled insulin pump. The attacker could alter dosage settings or disable alerts, putting the patient at serious risk. Such vulnerability issues are made possible due to the increased entry points from devices,[2] leaving those who use clinician-recommended technologies for care susceptible to attacks.

Additionally, connected medical devices can collect sensitive personal data, including protected health information and payment details. When this data is exposed outside of secure healthcare systems, it raises significant privacy concerns and opens the door for cyber attackers to exploit the information for malicious purposes. With a 2022 FBI Report[3] noting that 53% of devices have a known critical vulnerability, the growing interconnectedness of devices, and the quantity of cyberattacks impacting the healthcare industry, privacy concerns have become a significant issue.

The standard emphasizes a lifecycle approach to cybersecurity. It encourages engineers to embed security considerations from the earliest stages of product development through deployment and maintenance. Key components include:

- Authentication and authorization
- Data integrity and confidentiality
- Secure firmware updates
- Incident response planning

These are not theoretical concerns. In recent years, the U.S. Food and Drug Administration (FDA) has issued multiple safety communications[4] warning of cybersecurity vulnerabilities in medical devices.

For engineers, the challenge lies in balancing innovation with risk mitigation. Devices must be compact, power-efficient, and user-friendly, yet secure enough to withstand evolving cyber threats. This balancing act calls for a structured, standards-based approach.

IEEE 2621™: A STANDARDS-BASED SOLUTION

The IEEE Standards Association (IEEE SA) introduced the IEEE 2621™ series of standards[5] to address these challenges. This family of standards currently provides a framework for evaluating and certifying the cybersecurity of connected diabetes devices, with plans to expand to other device categories in the future.

IEEE 2621 aligns with national and international cybersecurity strategies, including the U.S. Food and Drug Administration's (FDA's) pre-market and post-market guidance, the National Institute of Standards and Technology (NIST) Cybersecurity Framework, and Section 524B of the Federal Food, Drug, and Cosmetic Act. It also incorporates best practices from IEC 80001-5-1 and AAMI TIR57.

The standard was developed through a collaborative effort involving manufacturers, clinicians, regulators, and cybersecurity experts. This multidisciplinary approach ensures that the standard addresses real-world threats while remaining practical for implementation.

INSIDE THE IEEE 2621 FRAMEWORK

IEEE 2621 provides functional requirements for wireless security evaluations. It outlines how to assess device resilience against common attack vectors such as unauthorized access, data interception, and firmware manipulation.

The standard emphasizes a lifecycle approach to cybersecurity. It encourages engineers to embed security considerations from the earliest stages of product development through deployment and maintenance. Key components include:

- *Authentication and authorization*: Ensuring only trusted users and systems can access the device
- *Data integrity and confidentiality*: Protecting patient data from tampering and unauthorized disclosure
- *Secure firmware updates*: Verifying the authenticity and integrity of software updates
- *Incident response planning*: Preparing for and mitigating the impact of security breaches.

By following IEEE 2621, engineers can design devices that not only meet regulatory expectations but also earn the trust of healthcare providers and patients.

CERTIFICATION: FROM COMPLIANCE TO COMPETITIVE ADVANTAGE

Cybersecurity readiness is much more than a simple value add. In today's regulatory environment, demonstrating readiness is an essential requirement. The IEEE Medical Device Cybersecurity Certification Program[6] developed under the IEEE Conformity Assessment Program (ICAP), offers manufacturers a clear, standardized path to validate the cybersecurity posture of their connected medical devices.

This program is built around the IEEE 2621™ series of standards defining functional requirements for wireless diabetes device security. However, the certification framework is designed to scale across a broader range of connected medical technologies.

The certification program follows a rigorous, multi-phase process designed to ensure both technical robustness and regulatory alignment

It provides a structured, third-party assessment process that aligns with FDA expectations and global regulatory trends.

THE CERTIFICATION PROCESS

The certification program follows a rigorous, multi-phase process designed to ensure both technical robustness and regulatory alignment:

1. *Pre-assessment and gap analysis*—Manufacturers begin with a pre-assessment phase, where ICAP experts evaluate the device's current cybersecurity controls against the IEEE 2621 standard. This step identifies gaps early in the process, allowing teams to make targeted improvements before formal testing begins.
2. *Formal testing and evaluation*—Accredited third-party laboratories conduct comprehensive testing using the IEEE 2621 Test Plan and Checklists. These tools ensure consistency, repeatability, and transparency in the evaluation process. Tests cover authentication, encryption, secure firmware updates, data integrity, and more.
3. *Standardized reporting and documentation*—Upon successful evaluation, manufacturers receive a Certification Report and Certificate of Conformity. These documents are formatted to support regulatory submissions, including FDA premarket filings. The standardized format reduces ambiguity and accelerates the review process.
4. *Ongoing surveillance and lifecycle support*—Certification is not a one-time event. The program supports ongoing surveillance and re-certification to ensure continued compliance as devices evolve through software updates or hardware revisions.

WHY CERTIFICATION MATTERS

For electrical engineers and compliance teams, the certification program offers several strategic advantages:

The IEEE Medical Device Cybersecurity Certification Program is more than a compliance tool; it's a catalyst for innovation.

- *Accelerated regulatory approval*—Certification aligns with FDA cybersecurity guidance and supports Section 524B of the Federal Food, Drug, and Cosmetic Act. This alignment can streamline pre-market submissions and reduce time-to-market.
- *Risk reduction and liability mitigation*—Manufacturers reduce the risk of post-market recalls, security breaches, and associated legal exposure by validating cybersecurity controls through an independent third party.
- *Market differentiation and trust*—Certification signals to healthcare providers, procurement officers, and patients that a device meets rigorous cybersecurity standards. It enhances brand credibility and can serve as a competitive differentiator in a crowded market.
- *Engineering efficiency*—The certification framework provides engineers with a clear roadmap for design and testing. It reduces guesswork, minimizes rework, and ensures that security is integrated from the ground up.
- *Global scalability*—The program is designed to harmonize with global standards, making it easier for manufacturers to meet compliance requirements across multiple jurisdictions.

THE ULTIMATE GOAL: SECURE, COMPLIANT INNOVATION

The IEEE Medical Device Cybersecurity Certification Program is more than a compliance tool; it's a catalyst for innovation. By embedding security into the design and validation process, manufacturers can bring safer, more innovative devices to market faster. For engineers, it offers a structured, standards-based approach to solving one of the most complex challenges in modern medical device development.

Choosing certification isn't just about checking a box. It's about building trust, protecting patients, and future-proofing your products in a rapidly evolving threat landscape.

STANDARDS INTEGRATION

IEEE 2621 doesn't exist in a vacuum. Other medical device-related workstreams from IEEE SA are also part of the conversation that integrate cybersecurity measures to protect patient privacy, ensure the integrity of medical data, and facilitate the interoperability of connected devices.

- The IEEE 11073 Standards Committee[7] collaborates with other global standards organizations to address the need for an openly defined, independent standard for controlling information exchange among connected personal health devices (PHDs) and the systems used to manage and control them (e.g., cell phones, personal computers, health gateways, etc.).

Connected medical devices are reshaping healthcare, but they also introduce new risks. Cybersecurity is now a critical component of product compliance.

- Another relevant IEEE SA activity is the Zero Trust Cybersecurity for Health Technology Tools, Services, and Devices Industry Connections Program.[8] It is a global community of technology stakeholders developing recommendations for a suite of new zero-trust network access (ZTNA) standards that integrate commercial and open-source products to showcase robust security features of Zero Trust Architecture (ZTA) when applied to enterprise IT use cases.
- The recently published IEEE/UL 2933™-2024, Standard for Clinical Internet of Things (IoT) Data and Device Interoperability with TIPPSS – Trust, Identity, Privacy, Protection, Safety, Security[9] is a TIPPSS framework for clinical (IoTs) data and device interoperability with healthcare systems including electronic health records (EHR), electronic medical records (EMR), other clinical IoT devices, in-hospital devices, and future devices and connected healthcare systems.

By harmonizing with these frameworks, IEEE 2621 supports global compliance efforts. Engineers working on products intended for international distribution can leverage the standard to meet diverse regulatory requirements without duplicating effort.

Connected medical devices are reshaping healthcare, but they also introduce new risks. Cybersecurity is now a critical component of product compliance, and engineers play a central role in securing the future of digital health.

LOOKING AHEAD

The IEEE 2621 series is just the beginning. Future iterations and other standards will expand to cover a broader range of medical devices, including cardiovascular implants, neurostimulators, and wearable diagnostics. The need for robust, scalable cybersecurity standards will only grow as the healthcare ecosystem becomes more interconnected, making way for entry points that have yet to be secured.

For engineers, this evolution presents both a challenge and an opportunity. Those who embrace cybersecurity as a core design principle will not only ensure compliance but will also drive innovation, protect patients, and strengthen their organization's competitive position.

Connected medical devices are reshaping healthcare, but they also introduce new risks. Cybersecurity is now a critical component of product compliance, and engineers play a central role in securing the future of digital health. By adopting standards like IEEE 2621, leveraging certification programs, and embedding security into every phase of development, engineers can build devices that are not only innovative but also safe, secure, and compliant.

The path forward is clear. Put security by design at the forefront. Then, align with global standards. Finally, lead the charge in building a safer, smarter, and patient-focused healthcare environment.

To learn more about the IEEE 2621 series of standards and other standards securing connected medical devices, please visit the IEEE Standards Association website[10] to see our latest news, including standards development and releases, as well as current and future webinars.

We always welcome volunteers to join our efforts, including our Medical Device Cybersecurity

Certification Program. Together, we can shape the future of medical devices and create a secure and innovative healthcare environment that can help patients on their health journey and providers in search of devices they can trust to provide care.

ENDNOTES

1. "Healthcare's Growing Threat Landscape," ISACA website, February 3, 2025.
2. "FDA reports potential cybersecurity risk with insulin pump system," American Hospital Association website, September 20, 2022.
3. "Private Industry Notification, Federal Bureau of Investigation, Cyber Division," September 12, 2022.
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5. "Addressing the Need for Protecting Cybersecurity in Connected Diabetes Devices," IEEE Standards Association, July 28, 2022.
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THE ENGINEER'S

TESTING HANDBOOK

Inside

Trust... But Verify

By Don MacArthur

Featured Experts

TRUST... BUT VERIFY

By Don MacArthur

Whether you're trying to get a product through safety approval from one of the many Nationally Recognized Testing Laboratory's (NTRLs) or having full compliance EMC testing performed at an accredited or non-accredited third-party laboratory or an internal in-house test facility, it's imperative for successful completion of the project that as a compliance engineer or technician leading a product certification effort, you trust but verify the work of these other entities. For the remainder of the article, we'll call these other entities "service providers."

WHAT IS MEANT BY THE PHRASE "TRUST BUT VERIFY"?

Trust means trusting those who may be doing work for us. We have already been through the process to qualify our service providers – we don't need to keep repeating that process each time we open a project with them. We know they are fully qualified and competent enough to carry out the activities we have asked them to do. We trust them to deliver the output of their work to the quality levels expected for our type of industry. Verify means following through with those we have entrusted to make sure they have completed the tasks we asked for, fully and accurately, before we sign off on completion.

"Trust but verify" is important for several reasons:

- ✓ You know the product better than the service provider.
- ✓ Service providers have other customers besides you.
- ✓ Service providers have limited resources – just like you.
- ✓ Service providers are always told schedule is a high priority.

WHY IS "TRUST BUT VERIFY" IMPORTANT?

"Trust but verify" is important for several reasons:

1. You know the product better than the service provider.

Those we've entrusted to do work for us don't know the product as well as we do and may miss important aspects that result in inaccurate testing or subpar certification or documentation.

One example is when EMC test personnel don't fully exercise the EUT during emissions testing, passing results are obtained, and the non-compliant product is allowed to ship. When in reality, had the EUT been exercised correctly, it would have failed, and a non-compliant product would not have been allowed to ship until a fix was in place. Whose fault is it that the product wasn't exercised correctly during EMC testing? One guess – it's not the service provider's fault. It's your fault!

Another example is having a safety agency review spacings on printed circuit board (PCB) and declare the board didn't meet the spacings requirements for the declared over-voltage category. The project was closed before you had a chance to react, and you had to accept a lesser over-voltage category rating for the product. Whose fault was it that the safety agency wasn't even reviewing the correct PCB layout in the first place? Yours!

2. Service providers have other customers besides you.

Whether it be internal or external service providers, service providers all have other customers they are working for besides just you. A lot of times, this means information overload for the service provider. Although your project might be important to them, if you're project is disorganized, it's going to take them a long time to sort through any issues given their heavy workload and all of the other projects they have going on at the time.

3. Service providers have limited resources – just like you.

Similar to item 2 above, service providers don't have extra resources to put into figuring out every aspect of testing your product. You need to take ownership and not assume you can just hand it over the wall for them to take care of for you. You need to be highly engaged, heavily involved, and be ready to supply information to your service provider before it's even asked for.

4. Service providers are always told schedule is a high priority.

Almost every project is a high-priority project if you're a service provider. One way to show service providers that your project is in fact, a high priority is to communicate with them on a regular basis as much as possible. Provide service providers with any information they're missing as soon as possible, hold regular status update meetings with them (at least weekly, more frequent if required), commit and deliver, and last but not least… continually "Trust but Verify."

REFERENCE AND FURTHER READING

1. Willink, J. & Babin, L., *Extreme Ownership – How U.S. Navy Seals Lead and Win*, St. Martin's Press, 2017.

Don MacArthur, known as "The Practical Engineer," is a Guest Contributor to *In Compliance Magazine* with over 30 years of hands-on experience in product development and global regulatory strategy. He has led engineering teams and certification efforts across defense systems, critical infrastructure, and industrial technologies. Don specializes in making complex compliance challenges accessible—translating standards, test methodologies, and conformance requirements into actionable engineering practices that drive certification success and product credibility.

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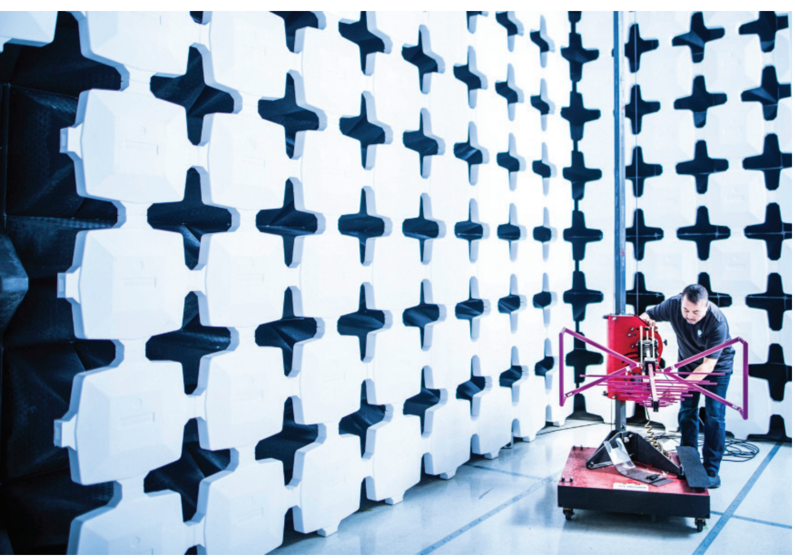

Element Materials Technology operates advanced battery testing labs offering comprehensive services for batteries used across industries like automotive, aerospace, consumer electronics, medical devices, and energy storage. These labs ensure batteries meet stringent safety, reliability, and performance standards, especially for lithium-ion, lithium-metal, and solid-state technologies.

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In addition, we offer reliability and ESD susceptibility testing of devices and systems using Human Body Model (HBM), Charged Device Model (CDM), and Machine Model (MM) simulators. These tests are vital for compliance assessments, failure analysis, and design verification, and are conducted in alignment with ANSI/ESD and JEDEC standards. This helps clients identify vulnerabilities early and validate robustness under realistic ESD scenarios.

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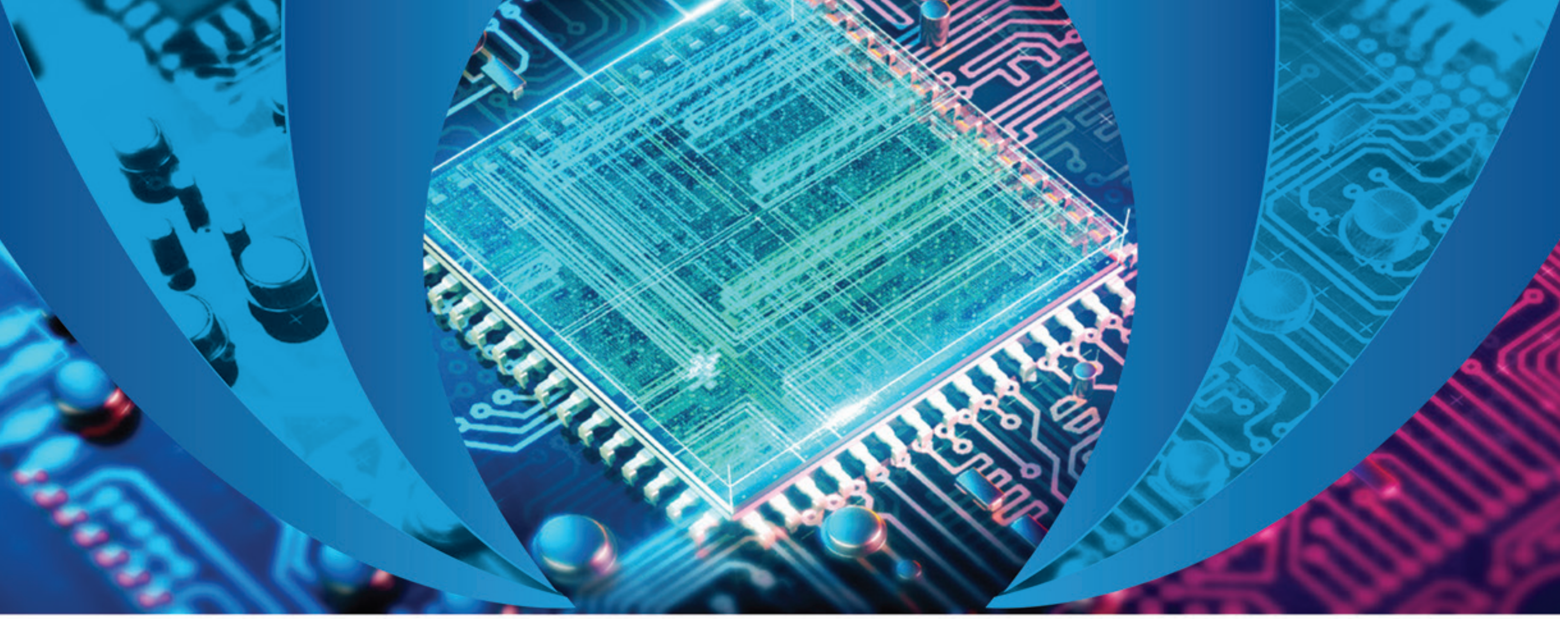

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To get the most out of your visit, we recommend coming with a clear understanding of your goals—whether it's qualification testing, R&D collaboration, or standards compliance—so we can tailor our discussion to your needs. Be ready to ask questions, explore the capabilities of our equipment and environment, and meet the team behind the testing. We encourage open dialogue, so bring samples, test challenges, or even just ideas you're exploring. It's not just a tour—it's a chance to align on solutions that support your product's success. *EOS/ESD Association Services, LLC*

To ensure a successful first visit, we recommend customers engage with us early in the planning process. Sharing detailed information about your device under test—including dimensions, frequency range, power requirements, and test objectives—allows us to prepare the appropriate range and resources in advance. We also encourage bringing any relevant documentation, such as test plans or compliance requirements, to streamline the process. Our team is here to support you every step of the way, so don't hesitate to ask questions or request a pre-visit consultation. The more we know ahead of time, the more efficient and productive your test session will be. *NSI-MI Technologies*

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"We've worked with NSI-MI across three antenna test campaigns for the satcom user terminal we're developing — and we're now preparing for a fourth major campaign. Their deep technical knowledge in antenna measurements (CATR and PNF) and testing standards has been invaluable throughout the process. NSI-MI is one of the very few facilities equipped to provide this level of service, offering world-class, highly accurate testing. Their responsiveness and attention to detail make them a standout in the field."

Amin Reda, Farcast

"I'd again like to like to express our sincere gratitude for the insightful knowledge sharing sessions you, Brennan, and Chuck recently conducted. We greatly appreciate you taking the time to share your expertise and best practices with us. We were particularly impressed by the professionalism, deep knowledge of the subject matter, and engaging communication style all of you fine folks at ESDA displayed. Your ability to explain complex concepts in a way that was easily understandable was invaluable to our team as we look forward to taking some big strides forward with our ESD program."

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SHIELDING TO PREVENT RADIATION

Part 3: Far-Field Shielding Effectiveness of a Solid Conducting Shield – Exact Solution

By Bogdan Adamczyk

This is the third of seven articles devoted to the topic of shielding to prevent electromagnetic wave radiation. The first article [1] discussed reflection and transmission of uniform plane waves at a normal boundary. The second article [2] addressed normal incidence of a uniform plane wave on a solid conducting shield with no apertures. The article concluded with the definition of shielding in the far field given by

$$(SE)_{dB} = 20\log_{10}\frac{|\hat{E}_i|}{|\hat{E}_t|} \tag{1}$$

This article presents the exact solution to Equation (1). The subsequent article will present the approximate solutions.

Figure 1 shows a conducting shield of thickness *t*, conductivity σ, permittivity ε, and permeability μ, surrounded on both sides by air (free space, and thus a perfect dielectric) [2]. Initially, there is no current density at the interfaces.

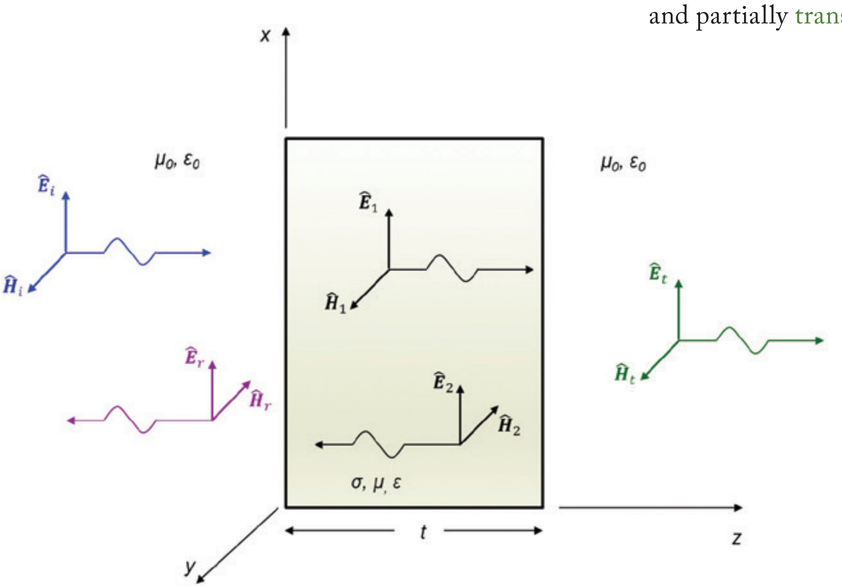

Figure 1: Uniform plane incident on a conducting shield

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A uniform plane wave is normally incident on its left interface. Uniformity assumption, together with normal incidence, means that the shield is in the far field of the radiation source.

The incident wave, upon arrival at the left most boundary $(\hat{\boldsymbol{E}}_i, \hat{\boldsymbol{H}}_i)$, will be partially reflected $(\hat{\boldsymbol{E}}_r, \hat{\boldsymbol{H}}_r)$ and partially transmitted $(\hat{\boldsymbol{E}}_1, \hat{\boldsymbol{H}}_1)$ through the shield.

The transmitted wave $(\hat{\boldsymbol{E}}_1, \hat{\boldsymbol{H}}_1)$, upon arrival at the right most boundary, will be partially reflected $(\hat{\boldsymbol{E}}_2, \hat{\boldsymbol{H}}_2)$ and partially transmitted $(\hat{\boldsymbol{E}}_t, \hat{\boldsymbol{H}}_t)$ through the shield.

The reflected wave ($\hat{\boldsymbol{E}}_2, \hat{\boldsymbol{H}}_2$) propagates back through the shield and strikes the first interface, incident from the right.

In [2], the incident wave is described by

$$\hat{\boldsymbol{E}}_i = \hat{E}_i e^{-j\beta_0 z}\boldsymbol{a}_x \tag{2a}$$

$$\hat{\boldsymbol{H}}_i = \frac{\hat{E}_i}{\eta_0} e^{-j\beta_0 z}\boldsymbol{a}_y \tag{2b}$$

The reflected wave is described by

$$\hat{\boldsymbol{E}}_r = \hat{E}_r e^{j\beta_0 z}\boldsymbol{a}_x \tag{3a}$$

$$\hat{\boldsymbol{H}}_r = -\frac{\hat{E}_r}{\eta_0} e^{j\beta_0 z}\boldsymbol{a}_y \tag{3b}$$

The wave transmitted through the left interface is described by

$$\hat{\boldsymbol{E}}_1 = \hat{E}_1 e^{-\hat{\gamma} z}\boldsymbol{a}_x \tag{4a}$$

$$\hat{\boldsymbol{H}}_1 = -\frac{\hat{E}_r}{\hat{\eta}} e^{-\hat{\gamma} z}\boldsymbol{a}_y \tag{4b}$$

The wave reflected at the right interface is described by

$$\hat{\boldsymbol{E}}_2 = \hat{E}_2 e^{\hat{\gamma} z}\boldsymbol{a}_x \tag{5a}$$

$$\hat{\boldsymbol{H}}_2 = -\frac{\hat{E}_2}{\hat{\eta}} e^{\hat{\gamma} z}\boldsymbol{a}_y \tag{5b}$$

Finally, the transmitted wave through the right interface is described by

$$\hat{\boldsymbol{E}}_t = \hat{E}_t e^{-j\beta_0 z}\boldsymbol{a}_x \tag{6a}$$

$$\hat{\boldsymbol{H}}_t = \frac{\hat{E}_t}{\eta_0} e^{-j\beta_0 z}\boldsymbol{a}_y \tag{6b}$$

The magnitude of incident field, $\hat{\boldsymbol{E}}_i$, is assumed to be known. In order to determine the magnitude of the transmitted field, $\hat{\boldsymbol{E}}_t$, we need to determine the magnitudes of the remaining waves, $\hat{\boldsymbol{E}}_r$, $\hat{\boldsymbol{E}}_1$, $\hat{\boldsymbol{E}}_2$. Thus, we need four equations in four unknowns. These are generated by enforcing the boundary conditions on the field vectors at the two boundaries $z = 0$ and $z = t$.

Continuity condition of the tangential components of the electric fields at the left interface produces [3]

$$\hat{\boldsymbol{E}}_i\Big|_{z=0} + \hat{\boldsymbol{E}}_r\Big|_{z=0} = \hat{\boldsymbol{E}}_1\Big|_{z=0} + \hat{\boldsymbol{E}}_2\Big|_{z=0} \tag{7}$$

or

$$\hat{E}_i e^{-j\beta_0 z}\boldsymbol{a}_x\Big|_{z=0} + \hat{E}_r e^{j\beta_0 z}\boldsymbol{a}_x\Big|_{z=0} = \hat{E}_1 e^{-\hat{\gamma} z}\boldsymbol{a}_x\Big|_{z=0} + \hat{E}_2 e^{\hat{\gamma} z}\boldsymbol{a}_x\Big|_{z=0} \tag{8}$$

leading to

$$\hat{E}_i + \hat{E}_r = \hat{E}_1 + \hat{E}_2 \tag{9}$$

Continuity condition of the tangential components of the magnetic fields at the left interface produces

$$\hat{\boldsymbol{H}}_i\Big|_{z=0} + \hat{\boldsymbol{H}}_r\Big|_{z=0} = \hat{\boldsymbol{H}}_1\Big|_{z=0} + \hat{\boldsymbol{H}}_2\Big|_{z=0} \tag{10}$$

or

$$\frac{\hat{E}_i}{\eta_0} e^{-j\beta_0 z}\boldsymbol{a}_y\Big|_{z=0} - \frac{\hat{E}_r}{\eta_0} e^{j\beta_0 z}\boldsymbol{a}_y\Big|_{z=0} = \frac{\hat{E}_1}{\hat{\eta}} e^{-\hat{\gamma} z}\boldsymbol{a}_y\Big|_{z=0} - \frac{\hat{E}_2}{\hat{\eta}} e^{\hat{\gamma} z}\boldsymbol{a}_y\Big|_{z=0} \tag{11}$$

leading to

$$\frac{\hat{E}_i}{\eta_0} - \frac{\hat{E}_r}{\eta_0} = \frac{\hat{E}_1}{\hat{\eta}} - \frac{\hat{E}_2}{\hat{\eta}} \tag{12}$$

Continuity condition of the tangential components of the electric fields at the right interface produces

$$\hat{\boldsymbol{E}}_1\Big|_{z=t} + \hat{\boldsymbol{E}}_2\Big|_{z=t} = \hat{\boldsymbol{E}}_t\Big|_{z=t} \tag{13}$$

or

$$\hat{E}_1 e^{-\hat{\gamma} z}\boldsymbol{a}_x\Big|_{z=t} + \hat{E}_2 e^{\hat{\gamma} z}\boldsymbol{a}_x\Big|_{z=t} = \hat{E}_t e^{-j\beta_0 z}\boldsymbol{a}_x\Big|_{z=t} \tag{14}$$

leading to

$$\hat{E}_1 e^{-\hat{\gamma} t} + \hat{E}_2 e^{\hat{\gamma} t} = \hat{E}_t e^{-j\beta_0 t} \tag{15}$$

Continuity condition of the tangential components of the magnetic fields at the right interface produces

$$\hat{\boldsymbol{H}}_1\Big|_{z=t} + \hat{\boldsymbol{H}}_2\Big|_{z=t} = \hat{\boldsymbol{H}}_t\Big|_{z=t} \tag{16}$$

or

$$\frac{\hat{E}_1}{\hat{\eta}} e^{-\hat{\gamma} z}\boldsymbol{a}_y\Big|_{z=t} - \frac{\hat{E}_2}{\hat{\eta}} e^{\hat{\gamma} z}\boldsymbol{a}_y\Big|_{z=t} = \frac{\hat{E}_t}{\eta_0} e^{-j\beta_0 z}\boldsymbol{a}_y\Big|_{z=t} \tag{17}$$

leading to

$$\frac{\hat{E}_1}{\hat{\eta}} e^{-\hat{\gamma} t} - \frac{\hat{E}_2}{\hat{\eta}} e^{\hat{\gamma} t} = \frac{\hat{E}_t}{\eta_0} e^{-j\beta_0 t} \tag{18}$$

Thus, we need to solve four equations: (9), (12), (15), and (18), repeated here

$$\hat{E}_i + \hat{E}_r = \hat{E}_1 + \hat{E}_2 \tag{19}$$

$$\frac{\hat{E}_i}{\eta_0} - \frac{\hat{E}_r}{\eta_0} = \frac{\hat{E}_1}{\hat{\eta}} - \frac{\hat{E}_2}{\hat{\eta}} \tag{20}$$

$$\hat{E}_1 e^{-\hat{\gamma} t} + \hat{E}_2 e^{\hat{\gamma} t} = \hat{E}_t e^{-j\beta_0 t} \tag{21}$$

$$\frac{\hat{E}_1}{\hat{\eta}} e^{-\hat{\gamma} t} - \frac{\hat{E}_2}{\hat{\eta}} e^{\hat{\gamma} t} = \frac{\hat{E}_t}{\eta_0} e^{-j\beta_0 t} \tag{22}$$

Towards this end, let us divide Eq. (21) by $\hat{\eta}$ to obtain

$$\frac{\hat{E}_1}{\hat{\eta}} e^{-\hat{\gamma} t} + \frac{\hat{E}_2}{\hat{\eta}} e^{\hat{\gamma} t} = \frac{\hat{E}_t}{\hat{\eta}} e^{-j\beta_0 t} \tag{23}$$

Adding Equations (22) and (23) gives

$$2\frac{\hat{E}_1}{\hat{\eta}} e^{-\hat{\gamma} t} = \hat{E}_t e^{-j\beta_0 t}\left(\frac{1}{\eta_0} + \frac{1}{\hat{\eta}}\right) \tag{24}$$

or

$$2\frac{\hat{E}_1}{\hat{\eta}} e^{-\hat{\gamma} t} = \hat{E}_t e^{-j\beta_0 t}\left(\frac{\hat{\eta}+\eta_0}{\eta_0\hat{\eta}}\right) \tag{25}$$

from which we obtain $\hat{\boldsymbol{E}}_1$ as

$$\hat{E}_1 = \hat{E}_t\left(\frac{\eta_0+\hat{\eta}}{2\eta_0}\right) e^{-j\beta_0 t} e^{\hat{\gamma} t} \tag{26}$$

Subtracting Eq. (22) from Eq. (23) gives

$$2\frac{\hat{E}_2}{\hat{\eta}} e^{\hat{\gamma} t} = \hat{E}_t e^{-j\beta_0 t}\left(\frac{1}{\hat{\eta}} - \frac{1}{\eta_0}\right) \tag{27}$$

or

$$2\frac{\hat{E}_2}{\hat{\eta}} e^{\hat{\gamma} t} = \hat{E}_t e^{-j\beta_0 t}\left(\frac{\eta_0-\hat{\eta}}{\eta_0\hat{\eta}}\right) \tag{28}$$

From which we obtain $\hat{\boldsymbol{E}}_2$ as

$$\hat{E}_2 = \hat{E}_t\left(\frac{\eta_0-\hat{\eta}}{2\eta_0}\right) e^{-j\beta_0 t} e^{-\hat{\gamma} t} \tag{29}$$

Next, let us divide Eq. (19) by η_0 to obtain

$$\frac{\hat{E}_i}{\eta_0} + \frac{\hat{E}_r}{\eta_0} = \frac{\hat{E}_1}{\eta_0} + \frac{\hat{E}_2}{\eta_0} \tag{30}$$

Adding Equations (20) and 30 gives

$$2\frac{\hat{E}_i}{\eta_0} = \hat{E}_1\left(\frac{1}{\eta_0}+\frac{1}{\hat{\eta}}\right) - \hat{E}_2\left(\frac{1}{\eta_0}-\frac{1}{\hat{\eta}}\right) \quad (31)$$

or

$$2\frac{\hat{E}_i}{\eta_0} = \hat{E}_1\left(\frac{\eta_0+\hat{\eta}}{\eta_0\hat{\eta}}\right) - \hat{E}_2\left(\frac{\eta_0-\hat{\eta}}{\eta_0\hat{\eta}}\right) \quad (32a)$$

from which we obtain $\hat{\boldsymbol{E}}_1$ as

$$\hat{E}_i = \hat{E}_1\left(\frac{\eta_0+\hat{\eta}}{2\hat{\eta}}\right) - \hat{E}_2\left(\frac{\eta_0-\hat{\eta}}{2\hat{\eta}}\right) \quad (32b)$$

Substituting for $\hat{\boldsymbol{E}}_1$ from Eq. (26) and $\hat{\boldsymbol{E}}_2$ from Eq. (29), we obtain

$$\hat{E}_i = \hat{E}_t\left(\frac{\eta_0+\hat{\eta}}{2\eta_0}\right)e^{-j\beta_0 t}e^{\hat{\gamma}t}\left(\frac{\eta_0+\hat{\eta}}{2\hat{\eta}}\right) - \hat{E}_t\left(\frac{\eta_0-\hat{\eta}}{2\eta_0}\right)e^{-j\beta_0 t}e^{-\hat{\gamma}t}\left(\frac{\eta_0-\hat{\eta}}{2\hat{\eta}}\right) \quad (33)$$

or

$$\frac{\hat{E}_i}{\hat{E}_t} = \left(\frac{\eta_0+\hat{\eta}}{2\eta_0}\right)e^{-j\beta_0 t}e^{\hat{\gamma}t}\left(\frac{\eta_0+\hat{\eta}}{2\hat{\eta}}\right) - \left(\frac{\eta_0-\hat{\eta}}{2\eta_0}\right)e^{-j\beta_0 t}e^{-\hat{\gamma}t}\left(\frac{\eta_0-\hat{\eta}}{2\hat{\eta}}\right) \quad (34)$$

or

$$\frac{\hat{E}_i}{\hat{E}_t} = \frac{\left(\eta_0+\hat{\eta}\right)^2}{4\eta_0\hat{\eta}}e^{-j\beta_0 t}e^{\hat{\gamma}t} - \frac{\left(\eta_0-\hat{\eta}\right)^2}{4\eta_0\hat{\eta}}e^{-j\beta_0 t}e^{-\hat{\gamma}t} \quad (35)$$

or

$$\frac{\hat{E}_i}{\hat{E}_t} = \frac{\left(\eta_0+\hat{\eta}\right)^2}{4\eta_0\hat{\eta}}\left[1-\frac{\left(\eta_0-\hat{\eta}\right)^2}{\left(\eta_0+\hat{\eta}\right)^2}e^{-2\hat{\gamma}t}\right]e^{-j\beta_0 t}e^{\hat{\gamma}t} \quad (36)$$

and thus

$$\frac{\hat{E}_i}{\hat{E}_t} = \frac{\left(\eta_0+\hat{\eta}\right)^2}{4\eta_0\hat{\eta}}\left[1-\left(\frac{\eta_0-\hat{\eta}}{\eta_0+\hat{\eta}}\right)^2 e^{-2\hat{\gamma}t}\right]e^{-j\beta_0 t}e^{\hat{\gamma}t} \quad (37)$$

Let us express the propagation constant $\hat{\boldsymbol{\gamma}}$ as

$$\hat{\gamma} = \alpha + j\beta \quad (38)$$

For a good conductor, the attenuation constant is related to skin depth by

$$\alpha = \frac{1}{\delta} \quad (39)$$

and thus the propagation constant becomes

$$\hat{\gamma} = \frac{1}{\delta} + j\beta \quad (40)$$

Using Eq. (40) in Eq. (37) gives

$$\frac{\hat{E}_i}{\hat{E}_t} = \frac{\left(\eta_0+\hat{\eta}\right)^2}{4\eta_0\hat{\eta}}\left[1-\left(\frac{\eta_0-\hat{\eta}}{\eta_0+\hat{\eta}}\right)^2 e^{-2\left(\frac{1}{\delta}+j\beta\right)t}\right]e^{-j\beta_0 t}e^{\left(\frac{1}{\delta}+j\beta\right)t} \quad (41)$$

and the shielding effectiveness becomes

$$SE = \frac{\hat{E}_i}{\hat{E}_t} = \frac{\left(\eta_0+\hat{\eta}\right)^2}{4\eta_0\hat{\eta}}\left[1-\left(\frac{\eta_0-\hat{\eta}}{\eta_0+\hat{\eta}}\right)^2 e^{-\frac{2t}{\delta}}e^{-j2\beta t}\right]e^{\frac{t}{\delta}}e^{j\beta t}e^{-j\beta_0 t} \quad (42)$$

The magnitude of the shielding effectiveness is

$$\left|\frac{\hat{E}_i}{\hat{E}_t}\right| = \left|\frac{\left(\eta_0+\hat{\eta}\right)^2}{4\eta_0\hat{\eta}}\left[1-\left(\frac{\eta_0-\hat{\eta}}{\eta_0+\hat{\eta}}\right)^2 e^{-\frac{2t}{\delta}}e^{-j2\beta t}\right]e^{\frac{t}{\delta}}e^{j\beta t}e^{-j\beta_0 t}\right| \quad (43)$$

or

$$\left|\frac{\hat{E}_i}{\hat{E}_t}\right| = \left|\frac{\left(\eta_0+\hat{\eta}\right)^2}{4\eta_0\hat{\eta}}\right|\left|\left[1-\left(\frac{\eta_0-\hat{\eta}}{\eta_0+\hat{\eta}}\right)^2 e^{-\frac{2t}{\delta}}e^{-j2\beta t}\right]e^{\frac{t}{\delta}}\right| \quad (44)$$

or

$$\left|\frac{\hat{E}_i}{\hat{E}_t}\right| = \left|\frac{\left(\eta_0+\hat{\eta}\right)^2}{4\eta_0\hat{\eta}}\right|e^{\frac{t}{\delta}}\left|1-\left(\frac{\eta_0-\hat{\eta}}{\eta_0+\hat{\eta}}\right)^2 e^{-\frac{2t}{\delta}}e^{-j2\beta t}\right| \quad (45)$$

It is convenient to express the shielding effectiveness in decibels

$$SE_{dB} = 20\log_{10}\left|\frac{\hat{E}_i}{\hat{E}_t}\right| \tag{46}$$

Then, utilizing Eq. (45), the *shielding effectiveness* in dB becomes

$$SE_{dB} = \underbrace{20\log_{10}\left|\frac{(\eta_0+\hat{\eta})^2}{4\eta_0\hat{\eta}}\right|}_{R_{dB}} + \underbrace{20\log_{10} e^{\frac{t}{\delta}}}_{A_{dB}} + \underbrace{20\log_{10}\left|1-\left(\frac{\eta_0-\hat{\eta}}{\eta_0+\hat{\eta}}\right)^2 e^{-\frac{2t}{\delta}}e^{-j2\beta t}\right|}_{M_{dB}} \tag{47}$$

or

$$SE_{dB} = R_{dB} + A_{dB} + M_{dB} \tag{48}$$

where R_{dB} is called the *reflection loss* and represents the portion of the incident field that is reflected at the shield interface. It is given by

$$R_{dB} = 20\log_{10}\left|\frac{(\eta_0+\hat{\eta})^2}{4\eta_0\hat{\eta}}\right| \tag{49}$$

A_{dB} is called the *absorption loss* and represents the portion of the incident field that crosses the shield surface and is attenuated as it travels through the shield. It is given by

$$A_{dB} = 20\log_{10} e^{\frac{t}{\delta}} \tag{50}$$

M_{dB} is called the *multiple-reflection loss* and represents the portion of the incident field that undergoes multiple reflections within the shield. It is given by

$$M_{dB} = 20\log_{10}\left|1-\left(\frac{\eta_0-\hat{\eta}}{\eta_0+\hat{\eta}}\right)^2 e^{-\frac{2t}{\delta}}e^{-j2\beta t}\right| \tag{51}$$

The reflection and absorption losses are positive numbers (in dB), while the multiple reflection loss is a negative number (in dB). It, therefore, reduces the shielding effectiveness.

The solution in Equation (42) or Equation (47) was obtained for the shield made of a good conductor under the assumption of normal incidence of the uniform wave, i.e., when the shield is in the far field of the radiation source. The solution in Equation (42) or (47) is often referred to as the *exact solution*.

In the next article, we will make some reasonable approximations that will greatly simplify this solution without any significant loss of accuracy.

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HOW TVS PROPERTIES AND PRINTED CIRCUIT BOARD DESIGN INFLUENCE PEAK VOLTAGE AND RESIDUAL CURRENT AT AN IC FOR USB-C SUPERSPEED DATA LINES

By Steffen Holland, Nima Lotfi, Martin Pilaski, Burkhard Laue, and Stefan Seider, Nexperia Germany GmbH

On behalf of EOS/ESD Association, Inc.

Founded in 1982, EOS/ESD Association, Inc. is a not for profit, professional organization, dedicated to education and furthering the technology Electrostatic Discharge (ESD) control and prevention. EOS/ESD Association, Inc. sponsors educational programs, develops ESD control and measurement standards, holds international technical symposiums, workshops, tutorials, and foster the exchange of technical information among its members and others.

The new connector standard USB-C includes both power delivery of current up to 20 V as well as SuperSpeed data lines. Integrated circuits (ICs) receiving signals on these data lines are very ESD sensitive. The Vbus pins in the USB-C connector are placed directly next to the SuperSpeed Tx and Rx pins, which poses the risk that the data pins can temporarily short to the supply voltage. The USB-C specification requires a mandatory AC coupling capacitor placed on the data lines in front of the IC when USB4 is used on the SuperSpeed lines. Thus, transient voltage suppressor (TVS) devices for system-level protection can be placed either behind the AC coupling capacitor or in front of it.

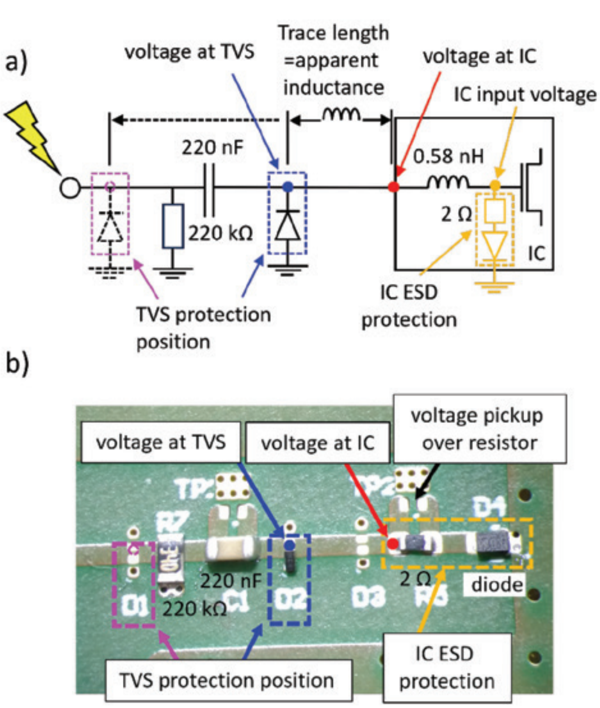

Figure 1: a) schematic representation of the application. b) real populated board with the smallest trace length distance between TVS and IC ESD replacement.

In this article, the effect of the TVS placement and the device properties on the IC is investigated. Special PCBs have been produced; one is shown in Figure 1. The distance between TVS and an IC replacement consisting of a 2-ohm resistor and a forward-bias diode is varied. The voltage at the IC can be measured. The IC residual current is determined by measuring the voltage drop across the resistor. Six different TVS protection devices have been chosen. All of them have a capacitance of less than 0.2 pF, which makes them suitable for the SuperSpeed application. Two of them are placed in front of the capacitor. However, to avoid turning on during a short to the power line, these TVS devices require a breakdown voltage larger than 20V. For a placement behind the capacitor, a high breakdown voltage is not needed. An overview of the TVS parameters is shown in Table 1.

TVS	position	$V_{trigger}$	V_{hold}	V_{peak}@16 A (trise =1 ns)
1	behind the capacitor	2.5 V	1.4 V	36.3 V
2		4.6 V	2.7 V	20.4 V
3		2 V	2 V	23.5 V
4		6.9 V	2.7 V	23 V
5	in front of the capacitor	31.2 V	1.6 V	55.8 V
6		39.2 V	20 V	159 V

Table 1: Overview of electrical parameters of chosen TVS devices.

Transmission Line Pulse (TLP) and Very-Fast-TLP measurements of the TVS devices have been performed with a rise time of 1 ns. The results are shown in Figure 2.

The Devices are chosen such that the different parameters have a wide range of values. TVS 1 to 4 have different trigger, peak, and holding voltages. A comparison allows for investigating the effect on the maximum input voltage and residual current into the IC. TVS 5 and 6 both have breakdown and trigger voltages larger than 20 V. The difference is that TVS 5 is a deep snapback device while TVS 6 does not snap back to below 20 V for all currents.

Figure 3a shows the residual current into the IC, and Figure 3b shows the peak voltage at the IC. While TVS 5 can strongly reduce the IC current after turning on, TVS 6 manages to divert only about 50% away from the IC. This is a direct effect of its high holding voltage. TVS 6 also exhibits a high peak voltage, which leads to a high peak IC input voltage. Even when the device turns on at 14 A, the peak IC input voltage is barely clamped. TVS 5 can limit the peak IC input voltage much more effectively. However, in comparison to the TVS 1 to 4 with their much lower trigger voltages, the absolute value is still higher. Compared to TVS 2, with the lowest peak IC input voltage, the value is 50% higher at 16 A. One can conclude that TVS devices with high trigger voltage, which can be placed in front of the AC coupling capacitor, would, in principle, be able to protect the capacitor. However, they are ill-suited for an

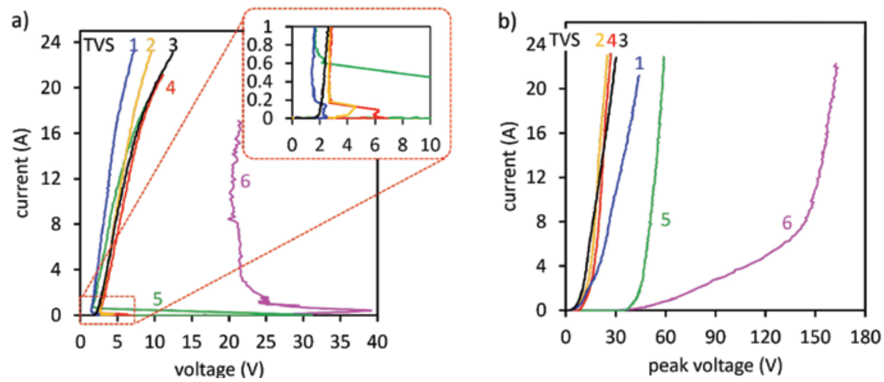

Figure 2 a) Quasi-static TLP curves of the TVS devices. The inset shows the curves at low currents b) Peak voltage of the components obtained by Very-Fast-TLP measurements.

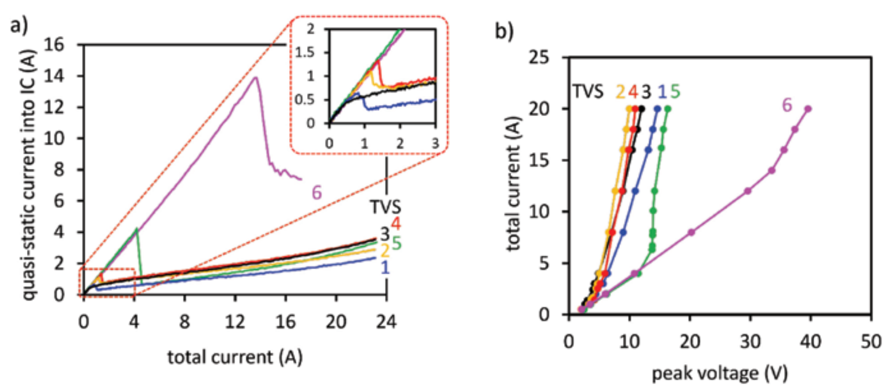

Figure 3: a) Residual current into the IC. The inset is a magnification of the low current range. b) Peak voltage at the IC input.

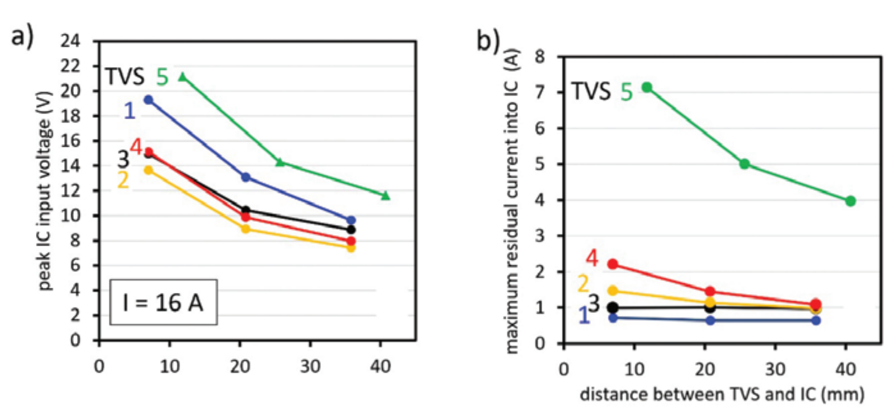

Figure 4: a) Peak IC input voltage and b) maximum residual current into the IC for different trace lengths.

The increasing prevalence of products with stacked dies necessitates defining and verifying design targets at low charging voltage levels.

effective system-level ESD protection of sensitive SuperSpeed ICs.

Surprisingly, for the trigger voltage range of TVS 1 to 4, the trigger voltage value has only a small effect on the peak IC input voltage. At high currents, it correlates with the peak voltage of the TVS device.

Figure 4 shows that a longer trace length between TVS protection and the IC replacement can significantly reduce the peak voltage at the IC as well as the maximum residual current into it.

In conclusion, placing a TVS protection in front of the AC-capacitor does not offer good IC protection. Low peak voltages at the IC can be achieved by lowering the peak voltage of the TVS protection and increasing the PCB trace length. The maximum residual current can be limited by reducing the trigger voltage of the TVS protection.

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CAN QUICK AND FREE SPICE SIMULATIONS HELP WITH EMI TROUBLESHOOTING?

By Dr. Min Zhang

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I recently heard an amusing quote from a colleague: "When it comes to EMC simulation, no one believes the results—except the person who did the simulation. When it comes to EMC testing, everyone believes the results—except the person who ran the test." Anyone who's worked in the field of EMC will probably smile knowingly at that.

Thanks to advancements in simulation tools, it's now possible to build fairly accurate models using 3D solvers to simulate conducted emissions, radiated emissions, surface currents, and more. But such simulation tools come at a price—not only in terms of expensive software licenses (which often puts them out of reach for small to medium-sized companies), but also the steep learning curve. Whoever runs these simulations needs to thoroughly understand the product and its circuitry—including parasitics, physical layout, and component behavior—and must also be skilled in building the simulation model itself. A good model can take weeks, even months, to develop. And after all that effort, how do you validate it? You still need test results to back it up.

As a practical engineer, I often lean more toward hands-on diagnostics. To paraphrase a well-worn saying: "My best simulation tool is my soldering iron." (Though in the EMC world, maybe we should say: "My best simulation tools are my near-field probes and current clamps.")

Still, is there a middle ground? Can simple, quick, and even free simulation tools help practical engineers in troubleshooting EMI?

After years of solving EMI issues in the field, I think the answer is yes—especially in specific scenarios. And I'm not talking about the heavy-duty full-wave simulators, but rather SPICE-based tools that are free to use and quick to learn.

Last year, Dr. Eric Bogatin presented a session on free power and signal integrity simulation tools at the EMC+SIPI symposium.[1] Coming from an electronics design background myself, I've always supported the use of free SPICE tools. My personal favorite is SIMetrix, though LTspice is just as capable.

So how do I use them in EMI work? Here are the three most common scenarios:

1. Filter design;
2. Testing assumptions and educating clients; and
3. Validating test results and informing next-gen designs.

Following are a few case studies illustrating each.

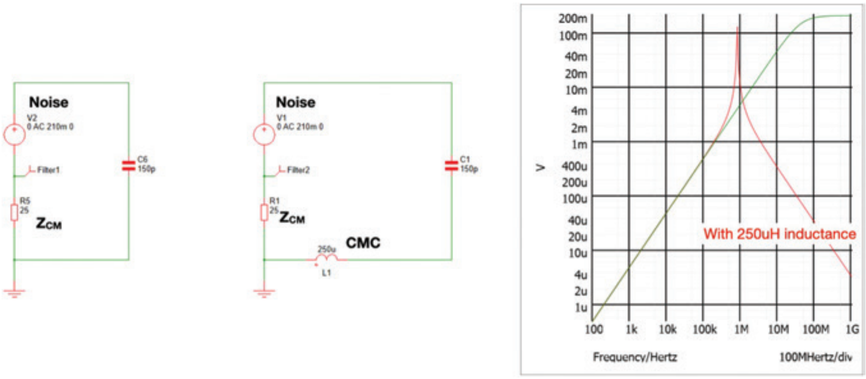

Figure 1: A quick and simple simulation tool for selecting a CMC

FILTER DESIGN WITH SIMPLE SPICE TOOLS

When building accurate filter simulations, the devil is in the details. Parasitics, coupling, and damping all play critical roles. As discussed in one of my papers,[2] it's not easy to get right. But when you're simply trying to determine how much impedance you need, a basic simulation might be enough.

In one example, I needed to estimate the required common-mode impedance/inductance for a small buck converter. I built a simple model in a few minutes: 25 ohms on the LISN side (for an automotive test setup, 25 ohm represents the common mode impedance[3]), with an estimated 150 pF parasitic capacitance between the converter and the test ground plane (you can simply use the math $C=\varepsilon_0\varepsilon_R A/h$, where A is the area of the PCB and h is the height of insulation support, in this case, 50mm). The model showed that with a 250 µH common-mode choke, noise attenuation would begin around 1 MHz.

Would I trust the model at higher frequencies? Absolutely not—it doesn't account for turn-to-turn winding capacitance or complex parasitics. But for the 1–100 MHz range, it gave me reasonable confidence in choosing a choke.

Figure 2 shows the actual measurement results. Noise reduction began around 1.5 MHz and had an effect up to about 100 MHz.

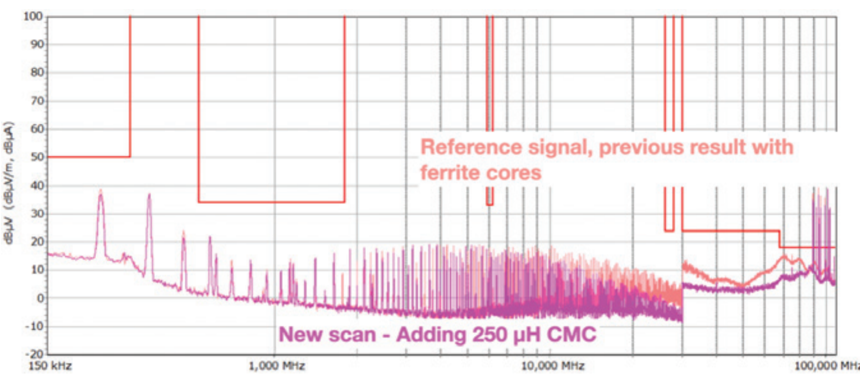

Figure 2: The selected CMC works from 1.5 MHz to about 100 MHz

SIMULATING EMC TEST SETUP ERRORS

Here's a rare but valuable case: sometimes, EMC test failures are due not to the product, but to incorrect test setups.

During one EFT/burst test, a product failed unexpectedly—even though it used a correctly installed, high-quality input filter from a reputable brand.

Upon investigation, I discovered a long ground lead had been connected to the wall in the shielded room (rather than a short ground lead bonded to the test ground plane).

To illustrate this to the team, I built a quick SPICE model. As shown in Figure 3, the simulation clearly demonstrated that a long ground lead (500nH inductance) could result in significant current spikes—validating our conclusion that the setup, not the product, was at fault. This was later confirmed during the test.

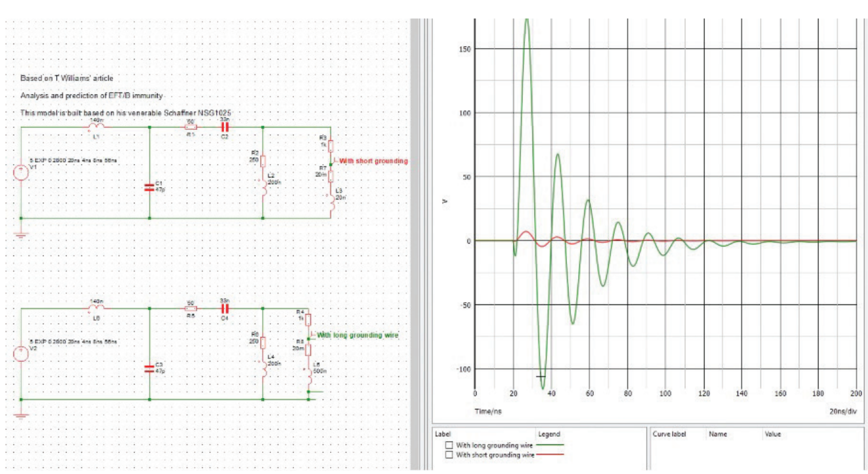

Figure 3: Simulation results showing overshoot and ringing in the current waveform due to the long ground lead

VALIDATING TEST RESULTS AND SUPPORTING THE NEXT DESIGN

In this case, a DC-DC converter using TO-247 packaged MOSFETs failed conducted emissions testing. The failure occurred in the 20 MHz range.

Oscilloscope measurements revealed overshoot and ringing during hard switching events. The resonant frequency of this ringing aligned perfectly with the EMC failure frequency.

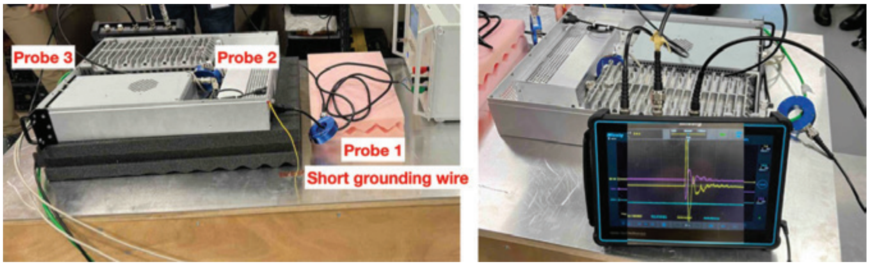

Figure 4: Lab results confirmed the findings— the yellow trace on the oscilloscope captured a current waveform that closely matched the simulation results.

We built a simple SPICE model of the switch node, incorporating parasitic inductance (primarily from the MOSFET leads). The simulation confirmed that the long legs of the TO-247 package were a major source of ringing.

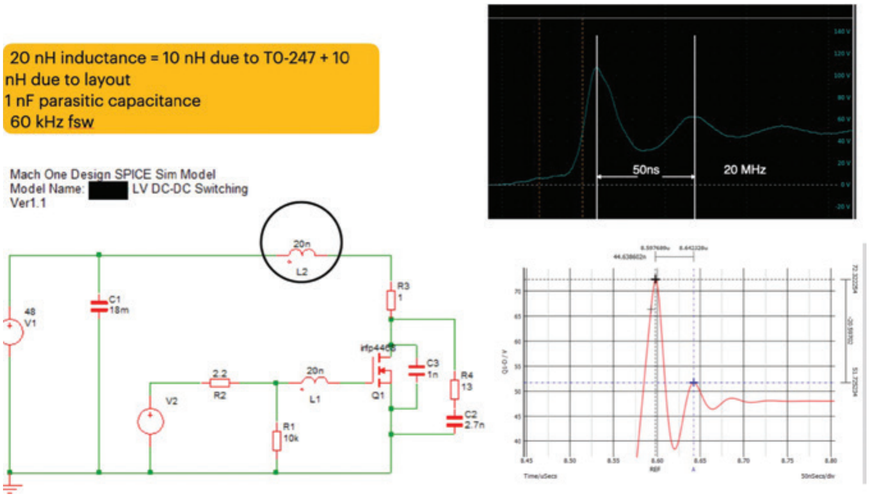

Figure 5: A simple simulation model on hard-switching devices

For the next-generation product, we strongly recommended switching to surface-mounted FETs to reduce inductive parasitics and improve EMC performance.

CONCLUSION

While full-blown EMC simulation tools are valuable, they're often impractical for fast troubleshooting or small teams. In contrast, quick, free SPICE-based simulations can be an efficient way to explore design ideas, test assumptions, and educate customers—without major investment.

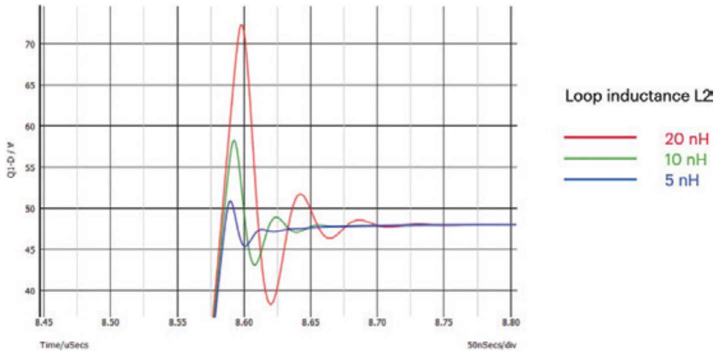

Figure 6: Simulation results show the importance of loop inductance caused by device package

They won't replace the lab bench, and they won't predict every nuance of your emissions profile. But when used wisely, they can be powerful tools in the hands of a practical engineer.

Sometimes, your best simulation tool really is your soldering iron and current probe—but your SPICE model might just be your second-best.

ENDNOTES

1. Eric Bogatin, "Five or More Favourite Free SIPI Tools," *Signal Integrity Journal* online, January 9, 2025.
2. Min Zhang, "A Simple and Effective SPICE-Based Simulation Model To Assist Your Filter Design," *Signal Integrity Journal* online, October 26, 2021.
3. Timothy Hegarty, "The Engineer's Guide To EMI in DC-DC Converters (Part 2): Noise Propagation and Filtering," *How2Power Today*, January 2018.

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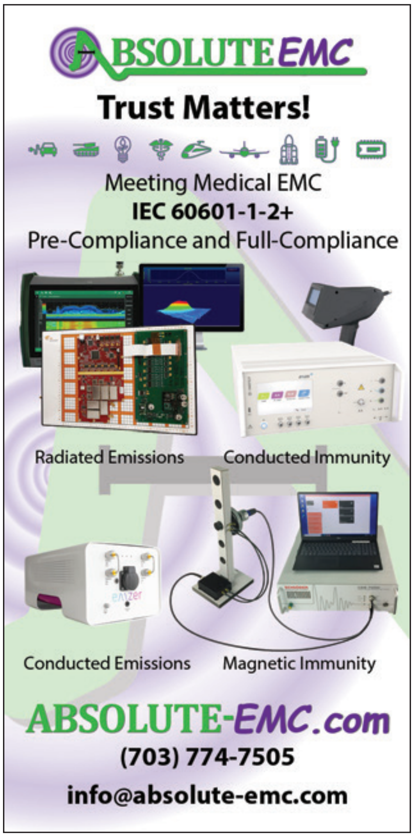

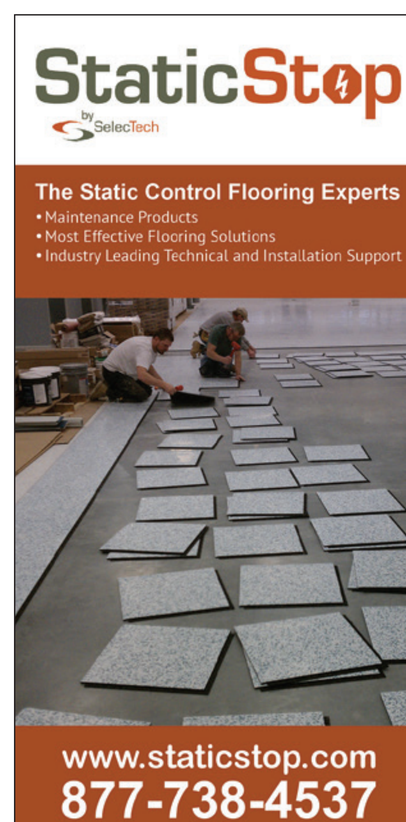

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