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CONTENTS

FDA Provides Update on ASCA Testing Laboratory Accreditation

The U.S. Food and Drug Administration (FDA) says it will release within the next 45 days an initial list of accredited testing laboratories under its Accreditation Scheme for Conformity Assessment (ASCA) pilot program.

The ASCA pilot accreditation program will allow accredited independent testing laboratories to assess medical devices for compliance with certain FDA-recognized standards. The establishment of the ASCA was mandated under the 2017 FDA Reauthorization Act and is expected to help facilitate a more efficient review process for certain types of medical devices.

According to an announcement posted on the FDA website, the list of accredited laboratories for the ASCA pilot program will be published by April 12th, with additional laboratories added to the list as they become accredited. Laboratories must complete a two-step process to become ASCA-accredited, which includes: 1) accreditation for conformity with the requirements of ISO/IEC 17025 and additional ASCA pilot program specifications; and 2) submit a request to the FDA to obtain ACSA-accreditation.

FCC Submits Report on Robocalls to Congress

The U.S. Federal Communications Commission (FCC) has released its annual report to Congress detailing consumer complaints and enforcement action in connection with illegal robocalls.

The report offers insight into trends related to complaints regarding robocall over a nearly six-year period from January 2015 through November 2020. Informal consumer complaints increased dramatically during the first four years covered by the report, from just over 200,000 in 2015 to more than 300,000 in 2018. But total consumer complaints then dropped significantly in the following two-year period, with 272,000 complaints filed in 2019 and an estimated 200,000 in 2020.

The drop in consumer complaints in 2019 and 2020 coincides with major actions undertaken by the FCC's Enforcement Bureau against robocall operators. The Commission issued three Notices of Apparent Liability for Forfeiture during that two-year period, with proposed forfeitures totaling nearly $250 million.

The Commission also issued two Forfeiture Orders in 2020, assessing penalties of nearly $50 million in connection with almost two and half million illegal telemarketing calls.

FDA Publishes Medical Device Shortage List

The U.S. Food and Drug Administration (FDA) has announced the publication of a medical device shortages list to help bring greater transparency regarding the availability of critical medical devices during the current pandemic.

Posted to the FDA's website last week, the device shortage list identifies more than 20 different medical devices for which limited supplies are currently available. The list includes a variety of ventilators and ventilator systems, specimen collection systems and devices, and personal protective equipment. The FDA says that it intends to continually update its device shortages list for the duration of the COVID-19 public health emergency.

The FDA cautions that the listing of a given device on the device shortages list does not necessarily indicate that patient care has been impacted by the shortage. However, it is publishing the device shortages list is part of its broader effort to "ensure that patients and health care providers have timely and continued access to high-quality medical devices."

The Pandemic Has Impacted the Semiconductor Chip Supply

The global COVID-19 pandemic has been with us now for a year, and it has had innumerable impacts on almost every aspect of our daily lives. But a recent posting on the CNBC website details an impact that few would have imagined possible.

It seems that the demand for computers and other electronic devices needed to support our new work-from-home population has increased to a record level. In fact, retail sales of electronics reportedly soared to a record $442 billion in 2020, with strong growth projected well into 2021.

But each of these devices requires a variety of semiconductor technologies, from central processing chips for computers to smaller, less expensive chips that control displays, peripherals, and communications devices. So, even with all-out efforts to increase semiconductor production, the overwhelming demand has forced suppliers to prioritize who gets the much-needed technology.

> The biggest impact of the chip shortage is falling on automobile manufacturers: GM, Ford, Chrysler, Honda, and Fiat, along with others. And Sony has said that the shortage of PlayStation 5 game consoles is due to the chip shortage.

According to the CNBC report, the biggest impact of the chip shortage is falling on automobile manufacturers. GM, Ford, Chrysler, Honda, and Fiat, along with others, have reportedly reduced or significantly slowed vehicle production due to the lack of semiconductor availability.

But the impact is also being felt closer to home (literally!), as Sony has said that the shortage of PlayStation 5 game consoles is due to the chip shortage.

RUSS DAVIS 1953-2020

It is with deep sadness that we are announcing the passing of Russ Davis, owner of JK Resources, who unexpectedly passed on December 24th, 2020.

Russ' journey to the EMC Community started in 1973, when he graduated from United Electronics Institute. He didn't know until 2008, that he had found the perfect community; the EMC Community. After several years of going to many of the events, Russ finally decided to start JK Resources in 2016 and continued to run JK Resources until his passing.

Russ loved this EMC community tremendously. He would always look forward to every single show, so he could catch up with old friends, make new ones, and talk about the ideas he was most passionate about. Russ had great pride in this community and everyone involved within it.

Russ was a man of dignity and was always prepared to lend a helping hand to anyone in need. His positive attitude, big smile, and kind heart will forever be missed.

Rosenworcel Designated Acting Chair of FCC

The Biden administration has designated Jessica Rosenworcel, long-time Commissioner of the U.S. Federal Communications Commission (FCC), to the position of Acting FCC chair.

Rosenworcel, who joined the FCC in May 2012, takes the helm of the agency following the departure of Ajit Pai, the former FCC chair. Pai resigned his position effective on January 20th, the day of President Joseph Biden's inauguration as President of the U.S.

According to a press release issued by the FCC, Rosenworcel has been a leader on spectrum policy and has continually sought ways to support the broad range of wireless services. She has also championed the work of women who have impacted digital life in the 21st Century through her podcast series "Broadband Conversations."

Prior to her time at the FCC, Rosenworcel served as Senior Communications Counsel for the U.S. Senate Committee on Commerce, Science, and Transportation. Before that, Rosenworcel practiced communications law in Washington, D.C.

FDA Names First Acting Director of Medical Device Cybersecurity

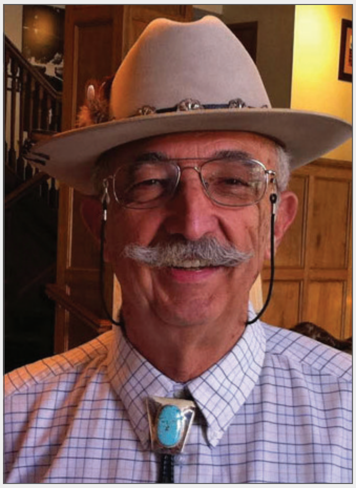
The U.S. Food and Drug Administration (FDA) has appointed its first Active Director of Medical Device Cybersecurity.

According to a posting on the website of Health ITSecurity, the agency has named Kevin Fu, an associate professor of electrical engineering and computer science at the University of Michigan, to the position. Fu is also the founder and chief scientist of the Archimedes Center for Medical Device Security.

The Medical Device Cybersecurity Director post falls under the scope of the FDA's Center for Devices and Radiological Health (CDRH) and has responsibility for leading the agency's efforts to ensure the safety and effectiveness of medical devices against cyber threats.

The Acting Director position at the FDA is intended as a 12-month position. During his tenure, Fu will reportedly continue his roles at the University of Michigan and the Archimedes Center.

JON BARTH 1937-2021



The engineering community lost a valuable asset this January with the passing of Jon Barth. His presence along with his contagious grand laughter will be very much missed at the ESDA symposiums and ESD work group meetings.

Jon was an inventor who always appreciated the challenge of an instrumentation problem. He started Barth Electronics in 1964 in his basement, with the design of a much needed high voltage, high speed attenuator for underground nuclear events at the Nevada Test site. After many successful years of designing and manufacturing high voltage instrumentation products for the national laboratories and the pulse power industry, Jon branched out into the ESD simulation and instrumentation industry. He leveraged his expertise with pulse power and manufactured the first commercial TLP machine for the ESD industry, which has been sold world-wide. His work has resulted in many patents in both the ESD and Pulse Power Industries. Authoring over a dozen ESD and Pulse Power Industry Papers, he was awarded the ESDA Industry Pioneer Award in 2006, and also the Nevada Exporter of the Year in 1991. He contributed as an active member in several ESDA Working Groups over the years, including Transmission Line Pulsing (TLP), Charge Device Model (CDM), System Level ESD, Transient Latch Up (TLU), and the Industry Council. He also served on the Technical Program Committee (TPC) for the ESDA mentoring authors. His willingness to bring to light technical deficiencies, and insight to address solutions, will be missed.

Jon often claimed he wasn't a "Device Guy," he was a "Measurement Guy." But he always took an active interest if a test of a customer's device showed it didn't perform as they expected. He would always help customers to fully understand the tests and measurements to ensure customer confidence in the test data.

Jon was that person who was always invested in helping people. He would thoughtfully help anyone figure out a practical and economical solution to make a difficult measurement. On many occasions, he could be heard advising a potential customer, "you don't need our product, you can simply do this or that, and you will get the data you need." He exuded integrity.

As a trusted voice, we looked to him as our mentor. Jon was always willing to listen and share his insightful knowledge. Never afraid to make a mistake, he always encouraged his employees to try new ways to improve the product or process. He was always interested in what could be learned from a failed attempt. He often said he succeeded because he had first tried and found all the options that didn't work. A philosophy, he attributed to a quote from Thomas Edison.

Quick to lend a helping hand to anyone in the community, he routinely helped students of all ages from kindergarten to grad students with an experiment or project. His inspiring quest for knowledge spanned a wide variety of topics, with special devotion and passion to science and engineering.

Jon was the "Measurement Guy," but one thing that Jon might have found hard, if not impossible, to measure is the impact he himself had on this industry. He was an amazing, caring, and sharing individual who left this world as he wanted, with a soldering iron in his hand.

Barth Electronics is committed to continuing Jon's tradition of providing quality state of the art ESD and Pulse Power products with the best in the industry service.

Jon's son, John Barth will lead the effort with the many other long time employees to continue to build on his legacy with new products and services.

DESIGN CONSIDERATIONS IN SPREAD SPECTRUM MODULATION FOR CISPR 25/CE TESTING

Discussing Factors that Influence the Design of Spread Spectrum Modulation for CISPR 25 Measurement

Christopher Semanson works at Renesas Electronics America Inc. as a Staff Power Systems Applications Engineer in Durham, NC supporting the design of PMICs and other power generation semiconductors in automotive applications in accordance with ISO 26262. He has five years previous experience in EMC Education at the University of Michigan, teaching EMC and Electronics with Mark Steffka. Semanson has a bachelor's degree in Electrical and Computer Engineering and a master's degree in Electrical Engineering from the University of Michigan Dearborn. He can be reached at christopher.semanson@renesas.com.

By Christopher Semanson

CISPR 25, CONDUCTED EMISSIONS, AND PWM CLOCK SOURCES

In automotive power electronics, there is one standard that has been either implemented or adopted as part of an OEM specification for interfacing and testing modules on a power bus, and that is a class of emissions limits defined in CISPR 25, "Vehicles, boats and internal combustion engines – Radio disturbance characteristics – Limits and methods of measurement for the protection of on-board receivers."

The purpose of CISPR 25 is to set limits to provide protection for receivers installed in a vehicle from disturbances produced by components/modules in the same vehicle, including switching power supplies, one of the biggest culprits of conducted disturbance. The standard covers frequency-dependent emission limits (in dB) and methods for test, including spectrum analyzer settings and test setup. It applies to electronics equipment and devices intended for use in vehicles or trailers.

To meet this design requirement, automotive power supply designers, including power management IC (PMIC) designers, have implemented a variety of different features in their products over the years. These features include:

- Slew rate control, where the designers can limit the slew rate of the switching signal and slow the pulse width modulation (PWM) switching speed; and
- Anti-phasic clock generation, where multiple modulators are driven out of phase with respect to each other.

Among these, however, is a more universal concept, spread spectrum clock generation, which has been widely adopted by digital designs and implemented on many modern-day power management devices. This allows the normally impulsive energy of a clock to be "spread" out over a wider frequency range, reducing the amplitude of the conducted emission generated by the clock. An example of this is shown in Figure 1.

Spread spectrum modulation applies to the PWM signal coming from what is commonly called the phase node used to drive either an external or internal switch. This switching on and off through an external filter is what gives high efficiency switch mode power supplies their allure. But it's also why they're the main source for conduced emissions (CE).

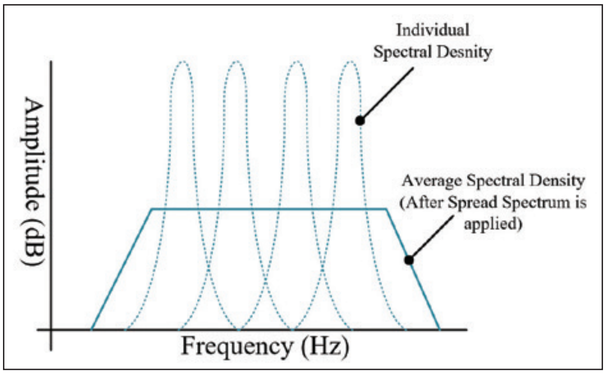


Figure 1: An example of how spread spectrum moves the energy of a clock signal

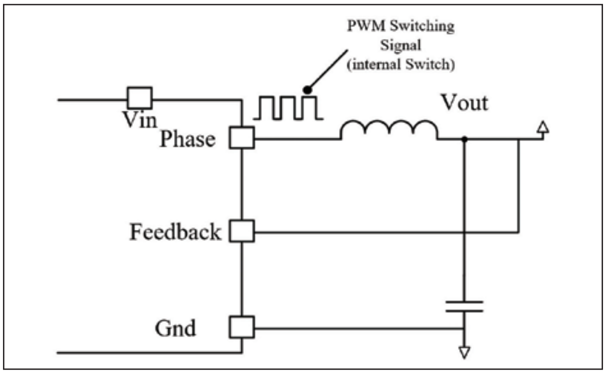


Figure 2: An example of a buck regulator including the switching signal

To understand how modulating the device turns the switch on and off and how it interacts with the standards, we'll start by giving an overview of CISPR 25 testing. Here, understanding of how the spectrum analyzer transforms a continuous time signal and represents it as a series of frequency spectra along with the proper limits is key to understanding the design of a proper spread spectrum generator. And, since this source of emissions can couple in non-predictable ways through unforeseen coupling paths, it's important to define a proper test setup so that repeatable results can be generated when evaluating schemes.

After identifying the important testing-related aspects outlined by this standard, we'll turn to the source of the problem itself, discuss the design goals of a spread spectrum scheme, and how it is used in a switch mode power supply. We'll focus on tunable factors commonly found as part of the design and outline the main design goals that go into modulating the clock.

The article will offer insight in how a spread spectrum function interacts with the measurement standard and how poor design can complicate matters. The goal is to demonstrate how spread spectrum can help the modulator meet emissions limits (or unwittingly make things worse!).

WHAT IS A SPECTRUM AND WHY DO WE MEASURE IT?

When designing or evaluating a design according to CISPR 25, understanding how a continuous time signal is represented in signal spectra is key in understanding how these two topics relate to each other. In theory, spectrum can be defined as a collection of sine waves superimposed on each other in the time domain and in the frequency domain, with each sine wave operating within its own fundamental frequency.

However, in reality, it's not that simple since we often have a DC signal with switching transients superimposed on top of it, as shown in Figure 3.

In both cases, both domains help characterize the design in different ways:

- The time domain helps with functional details such as overshoot, ringing, settling time, and undershoot and is more closely related to devices *functional* performance.
- The frequency domain helps with regulatory and integration details, which ensures that the strength of those transient signals is within a limit that doesn't impact neighboring electrical devices and signals. These features are more often referred to as *regulatory* performance.

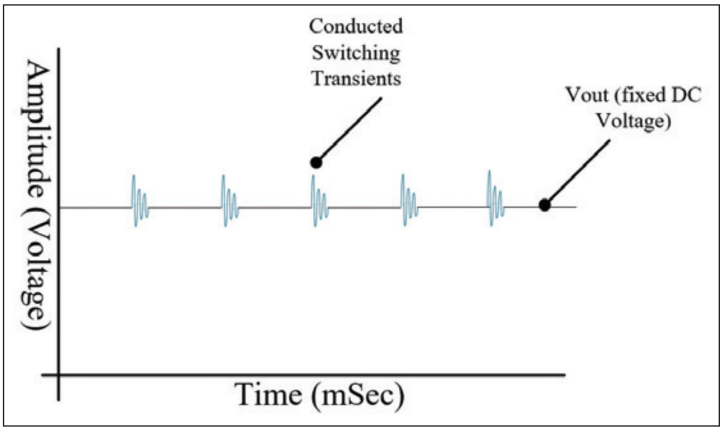


Figure 3: An example picture of a DC signal with a switching transient superimposed upon it

To do this, we employ a family of devices that allow us to analyze these continuous time signals and represent them in the frequency domain, classified as signal analyzers. These devices do their job through a various combination of analog and digital circuits. This family of devices covers spectrum analyzers, EMC analyzers, vector generators, and, sometimes, even the "FFT" function on an oscilloscope.

The operating principles of these devices are similar. But to keep things simple, the focus of this section will be a swept-tuned superheterodyne measurement device, a type of spectrum analyzer that sweeps and measures a range of frequencies higher than those used for audio communications. Our goal is to understand how the frequency impacts the measurement and how the measurement process measures and captures a continuous time signal to represent it as signal spectra.

Defining the Fundamentals of a Spectrum Analyzer

To understand why CISPR outlines not only the test setup but also the bandwidth settings, our attention needs to turn to the fundamental block diagram of how a spectrum analyzer functions. It is made up of

subsystems that limit the input, mix it with a local oscillator, and then various filters and conditioning circuits that aid in the display of the signal's spectra. An example block diagram is shown in Figure 4.

Next, we'll go into more detail for each of the subsystems.

Input Attention

The first subsystem is the input signal conditioning circuit. This subsystem is generally made of an attenuator and often a pre-selector filter. Its focus is to ensure that the signal enters the mixer at the optimum level for measurement. An example is shown in Figure 5.

By providing input attenuation, the device ensures that the signal is free from distortion by the spectrum analyzer by limiting it to a suitable input level for the downstream circuits, thereby providing physical protection and creating a suitable noise floor.

Mixing

Next is the mixer circuit. This subsystem, often made up of more than one mixing step, synchronizes the span of the screen to the portion of the signal spectra being measured. The basis of this circuit is where the device gets its name, heterodyne, which means to mix. By mixing the input signal spectra with a local oscillator, a new signal is created based on the local

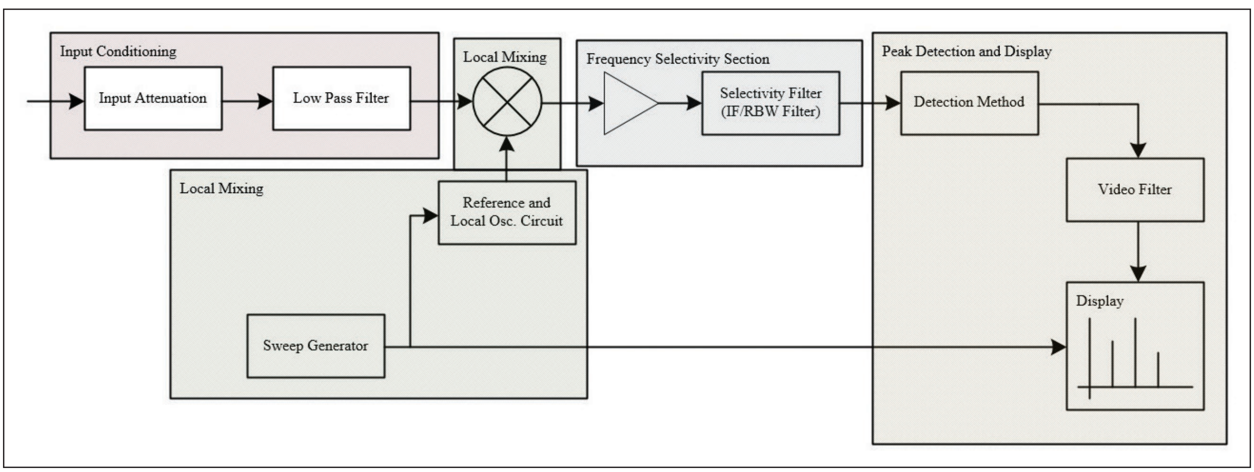


Figure 4: Block diagram of a super heterodyne swept tuned frequency spectrum analyzer

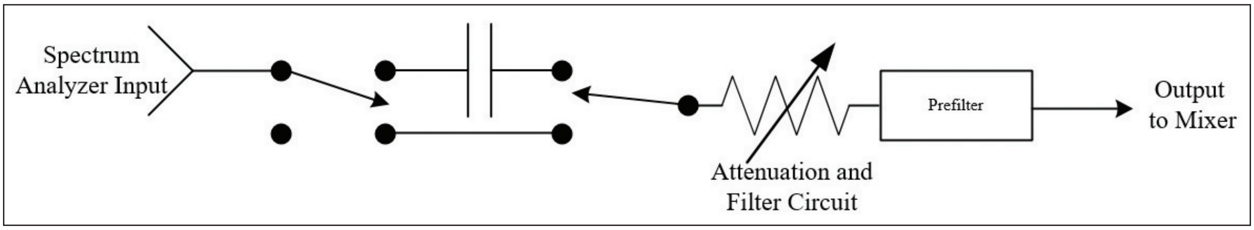


Figure 5: Example attenuation subsystem

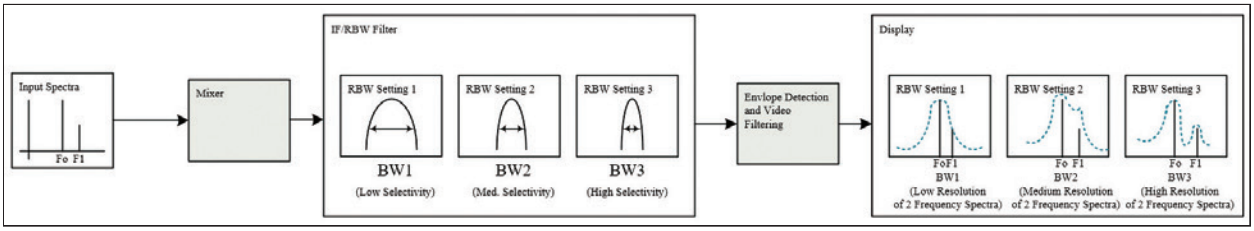


Figure 6: Example of how the RBW of a filter impacts the display/measurement

oscillator which includes not only the original signal but the harmonics which need to be filtered.

The Intermediate Filter (IF) and Gain

After mixing, we need to filter and adjust the positions of the signals on the display without impacting the signal level at the input mixer. As the IF gain level is changed, the value of the references changes in relation to that. Next is the filter itself, which exists to create resolvability between signals of unequal amplitude. The filter exists as a bandpass filter of configurable bandwidth called the resolution bandwidth (RBW) that determines selectability. An example of this discrimination is shown in Figure 6.

What was once an analog bandpass filter with poor selectivity and wide bandwidth now exists as a digital filter with the ability to decrease the bandwidth of the filter.

The Detector and Video Bandwidth Filter

After the input signal passes through the IF filter, it goes through a detector that outputs a signal, the amplitude of which is proportional to the power passing through the IF filter. This detector once existed as a pure analog part (modeled as a diode) which detected only the peaks. With the advent of digital spectrum analyzers, more examples include:

- Negative/positive peak
- Normal
- Average
- Quasi peak, or
- Sample

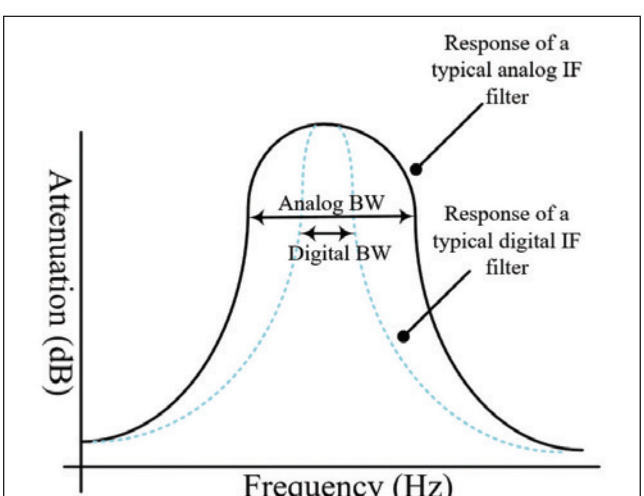
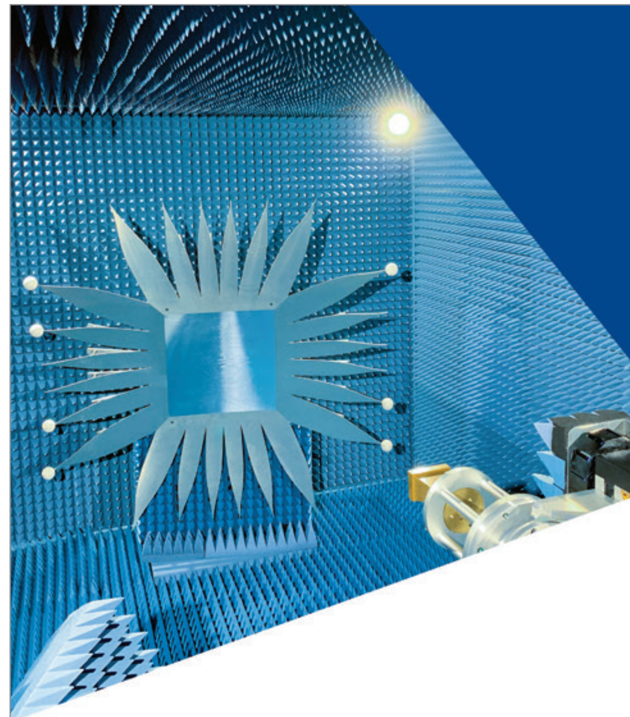


Figure 7: An analog filter with poor selectivity compared to a digital filter



The difference in how the detector operates is based on the specification being followed. For example, positive peak is often the default on most devices as it ensures that no sinusoid is missed in the display.

The difference in how the detector operates is based on the specification being followed. For example, positive peak is often the default on most devices as it ensures that no sinusoid is missed in the display (it displays every peak frequency scanned). But in comparison with other modes, it may not give a good representation of random noise in the system when testing broad band noise from a motor since it only detects the maximum value at each frequency and ignores the "randomness" of the brushed motor noise. The detection type is generally matched to the application.

Lastly, before the detector displays the output, the input signal is passed through another filter, the video filter. Unlike the IF, this filter is usually matched to the span and display settings of the analyzer. Its purpose is to smooth the noise inherent to the signal being measured, and its bandwidth is called the video bandwidth, or VBW. The narrower the bandwidth, the less noise there is in the output signal. But, once again, the narrower the bandwidth, the longer the sweep time and overall impact on the amplitude of the signal being measured.

Tunable Parameters

The main tunable parameters of a spectrum analyzer are:

- The input attenuation setting, which while often set can be tuned to create a more optimum baseline in which to view the signal spectra;
- The frequency span, which impacts both how much of the signal spectrum you're seeing and the sweep time;
- The bandwidths, which impact how much of the spectrum is being represented on the screen; and
- How the envelope of the signal is detected.

To find out how the shape of these filters relate to the representation of the signal, we'll investigate how their bandwidth can impact the amplitude, and how it relates to CISPR requirements.

Video and Resolution Bandwidth and How It Relates to CISPR

The primary purpose of the RBW filters that exist after the mixing stage is to resolve adjacent signals of different frequency representations. The key term here is "adjacent signals," which plays an important role in addressing how the spectrum analyzer can be configured to properly represent the signal. We do that by first defining two different types of signals usually measured in a conducted emissions test:

- Narrow banded signals, usually caused by a grouping of signals inside of a narrow band of frequency spectra. Usually, these signals have a defined frequency and well-defined amplitude and change based on the defined bandwidth; and
- Broad banded signals, signals which show no change in maximum defined amplitude with changes in spectrum analyzer bandwidth.

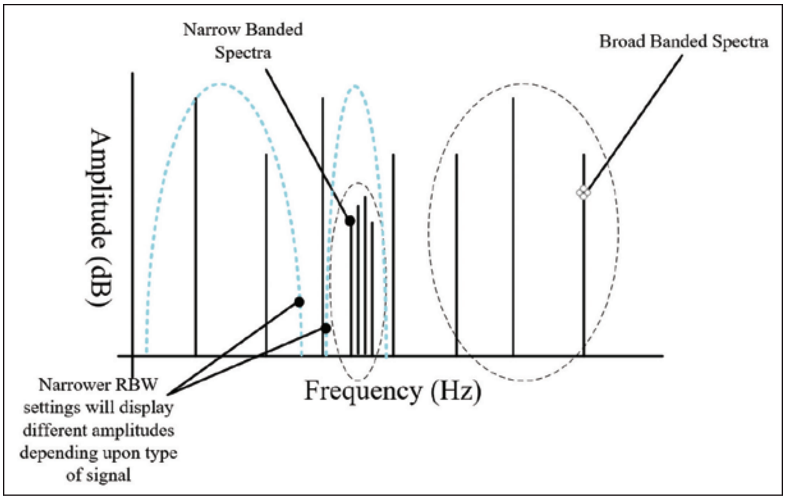


Figure 8: Difference between broad band, and narrow banded signals

As the RBW filter is modified and the signal amplitude changes, the definition of the signal becomes more apparent. An example of the selectivity is shown in Figure 9.

By decreasing the bandwidth of the filter, we decrease the amount of signal being measured. This helps in isolating narrow banded signals, but the result might not be representative of broad banded signals.

A clock can fit into the definition of narrow banded signals when the fundamental frequency has jitter or is frequency modulated. In general, they are usually represented as a fundamental frequency and their associated harmonics at multiples of the base frequency that scale based on the fundamental frequency. For example, if we have a base clock switching at 2MHz, we'll see a spectral line at 2MHz, 4MHz, 6MHz, and so on, with their spacing being a function of the base switching frequency. Great care must be taken when considering the RBW settings for the following reasons:

- If set too high, the RBW will not be able to distinguish between adjacent signals (or harmonics), and you may end up capturing more than the frequency of interest.
- If set too low, the RBW will not be able to capture all the spectral lines of interest.

Ideally, you want to set the RBW as narrow as possible to capture the frequency of interest. However, doing so comes with an increase in sweep time. Luckily, to accurately measure the signal from the LISN, CISPR 25 defines the bandwidth and scan time based on the frequency range of interest depending upon the detection method. Some examples are shown in Table 1 on page 18.

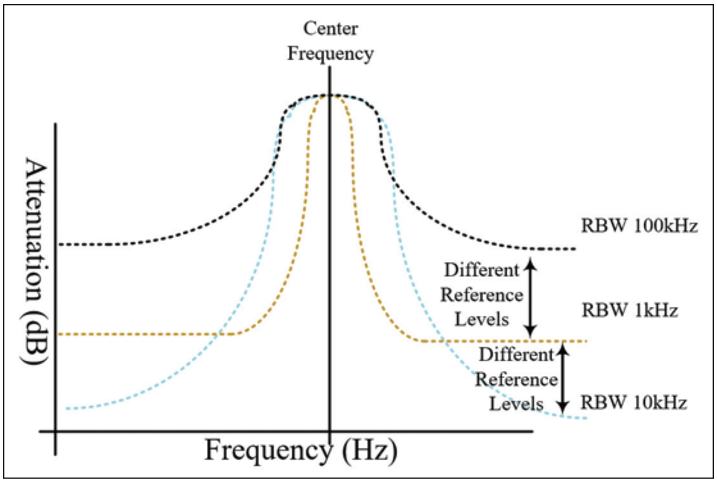


Figure 9: Example of how RBW/IF filter can increase or decrease selectivity

It's important to take note that the full definition of a Class 5 measurement in CISPR 25 defines the following as a function of the frequency range or spanning range:

- A noise floor, which is impacted by the input attenuator of the analyzer;
- The resolution bandwidth; and
- The sweep time, a direct function of how much the span is.

Next, we'll discuss how this relates to proper design of a spread spectrum generator when choosing exactly how close the frequencies hop between and the shape of the slope of the clock.

WHAT DOES IT MEAN TO DESIGN A SPREAD SPECTRUM MODULATOR?

Spread spectrum techniques are abundant in clocking and are used to deliberately spread the frequency of the clock in the frequency domain, resulting in a wider

Frequency Range	Detection Type					
	Peak		Average		Quasi Peak	
	RBW	Scan Time	RBW	Scan Time	RBW	Scan Time
.15MHz-30MHz (AM Broadcast)	9/10kHz	10 s/MHz	9KHz	200 s/MHz	9/10kHz	10 s/MHz
76MHz-108MHz (FM Broadcast)	100/120 KHz	100 ms/MHz	120 KHz	20 s/MHz	100/120KHz	100ms/MHz

Table 1: A excerpt of emissions requirements from CISPR 25

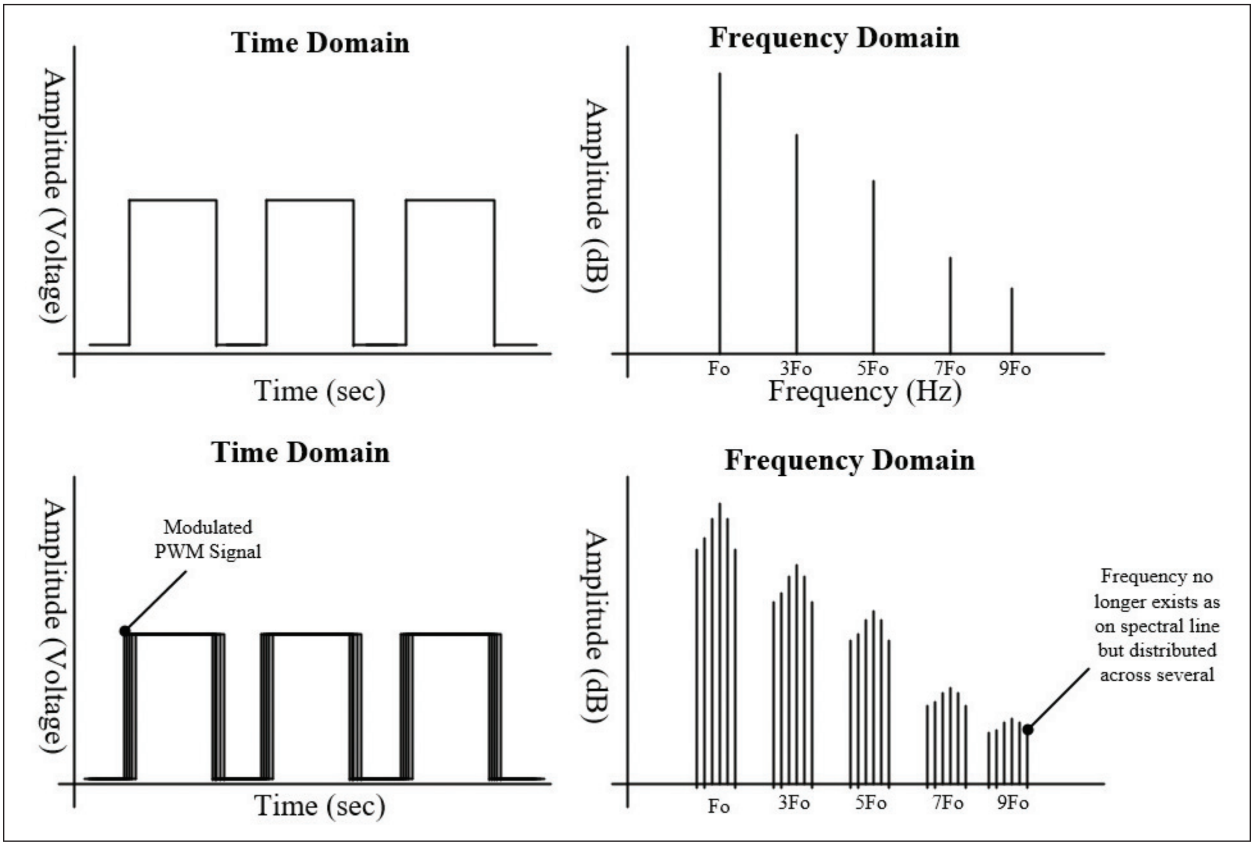


Figure 10: Time domain vs. frequency domain of a modulated and unmodulated PWM clock

bandwidth with lower amplitude. Its main advantage to the system is that, by reducing the source of conducted emissions (i.e., the PWM clock), the peak emission is spread over a larger frequency range.

A physical demonstration of this modulation scheme is demonstrated in Figure 10. It compares the measurements before and after clocking modulation is turned on, represented in both the time and frequency domain.

We characterize the amount of modulation as a percentage of the targeted clock signal, typically measured in the kHz range. For example, if we take a 1MHz switching signal and modulate it by 1%, the frequency range would dither between 990kHz and 1.05MHz, resulting in a signal spectrum that no longer spends all its time at the fundamental frequency of 1MHz. In a power regulator, the two ways to set this dithering are frequently either:

- An external resistor which, when coupled with a current source, produces a voltage that translates to a percentage; or,
- In more advanced applications, via one-time programmable fuse options where both the frequency dithering percentage and shape of the clock is chosen internally to the device.

Next, we'll discuss design decisions as they pertain to a spread spectrum modulation scheme and then relate it to the RBW of the spectrum analyzer.

Design Decisions in Spread Spectrum Modulation

At its core, the spread spectrum modulator is a frequency modulation scheme that controls the jitter of the PWM in the time domain by algorithmically bouncing around the fundamental clock frequency. The result of this scheme in the frequency domain is a spectral line that no longer exists at just one frequency but instead at multiple, similar frequencies.

The result is a reduction in both the peak and overall average energy at the targeted switching frequency (Fsw). The diagram shown in Figure 11 demonstrates an algorithmic flow chart representative of a frequency hopping spread spectrum generator.

And while the most optimal dither is of debate, it is often targeted to the application, there exists a combination of parameters that can be tuned for a specific situation. The options pertain specifically to the dithering or jumping of the fundamental PWM frequency. Some popular design decisions that impact the dithering amount are:

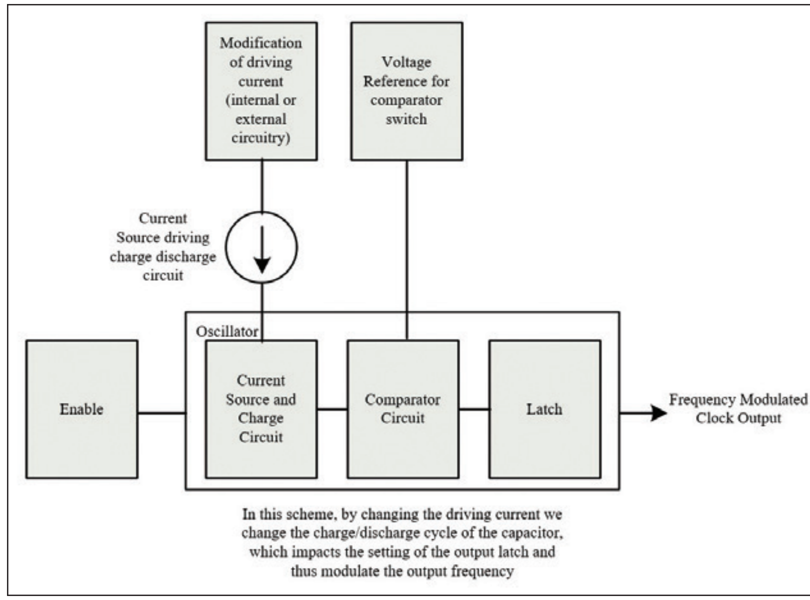


Figure 11: An Example of a Spread spectrum generator block diagram

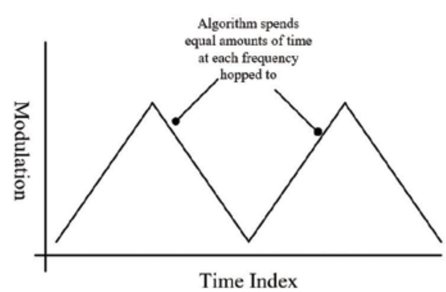
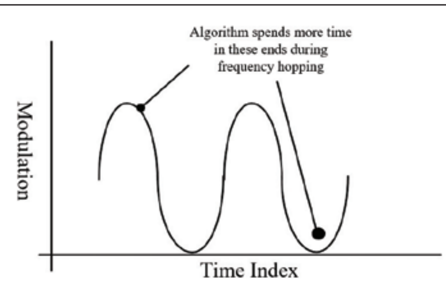
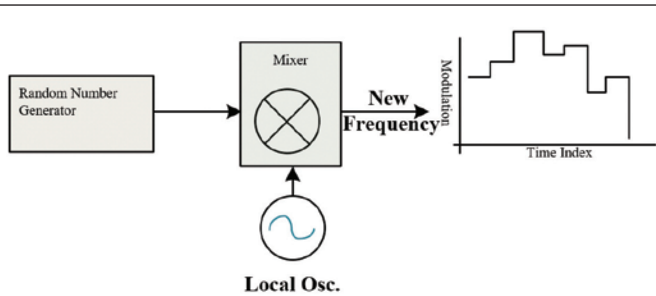
Waveform Representation	**Name**	**Description**
	Triangular/Linear	Circuit linearly cycles through available frequencies, spending equal amounts of time at each frequency.
	Pseudo-random	The clock signal jumps around with a dwell time and frequency jump denoted by a random number generator.
	Sinusoidal	Circuit spends majority of the time near the ends of the targeted frequency, less in the middle.

Table 2: Examples of types of spread spectrum modulation schemes

- Controlling the frequency being hopped through pseudo-random frequency hopping or a specific modulation scheme. Instead of linearly running through the list of allowable frequencies, it hops between them with a random number generator, according to an algorithm or external circuit.

Table 2 demonstrates this concept with different schemes. This is basically understood as how long the clock spends at each frequency between the minimum and maximum.

- Controlling the dwell time, or how long the clock frequency spends at each frequency that is hopped to. This can also be random. In general, a smaller dwell time means a more rapid repeating of each frequency hop, but too long or too short of a dwell time impacts the average noise measurement.
- Controlling the spacing of the frequency hops. We want to make sure that, as the frequency dithers between the maximum allowable limits, the hops are far enough apart to be distinguishable. The important thing to note here is that the dithering can be either continuous or digital (an example is shown in Figure 12).

And, in a complex mixed-signal design, there will often be more than one implementation selectable via fuse or register that will allow for evaluation of the best implementation. The overall goal is to reduce both the:

- Peak amplitude of the fundamental frequency of the PWM clock; and
- The average power associated with each harmonic.

An example is shown in Figure 12 of a before and after spread spectrum, with both the average and peak signals taken.

Now that we've introduced how RBW impacts the measurement taken and how a spread spectrum modulation scheme impacts the frequency spectrum, we can put the two together to understand how they impact the measurement and obtain some design goals.

SPREAD SPECTRUM AND HOW IT INTERACTS WITH THE RBW AND CISPR

As we covered in the previous section, we're able to shape the signal spectra of the PWM signal by choosing the modulation scheme. This results in the amplitude of each of these spectral lines being impacted by choosing parameters that impact the dwell time and the frequency hopping. If the RBW parameter is overlaid on top of what the signal spectra looks like, some design goals are quickly realized based on the interaction of measurement and function. Specifically:

1. The tightness of the SS frequency hopping must not be so close together that two or more frequencies fit under the same RBW filter. By creating a narrow-banded clock spread, the result would be an increase of signal spectra at that frequency instead of a decrease.
2. The bandwidth of the dithered signal as it relates to the RBW is extremely important as it relates

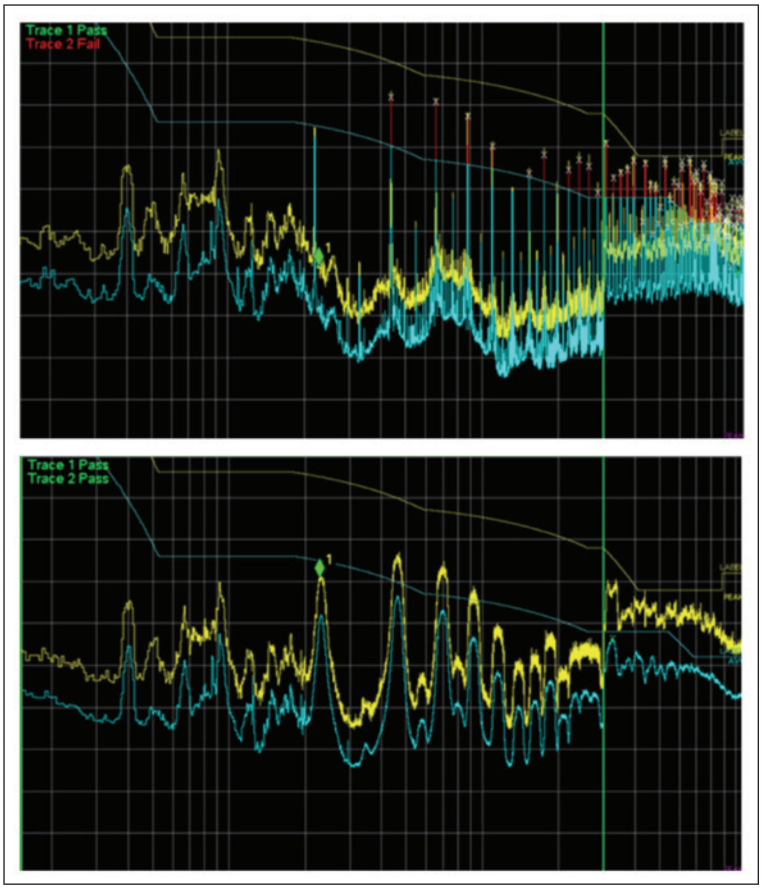


Figure 12: Data showing before and after spread spectrum is turned on

to the spread of the energy being measured. If the dithered signal has a very narrow bandwidth, the peak amplitude is likely to be high. Spreading the bandwidth to take advantage of the full RBW of the spectrum analyzer will reduce the average of the harmonic being measured at each RBW interval.

By following these design goals, we can thereby control the peak and average emissions. This also means that a spread spectrum generator that produces very tight discrete frequency bands may not always be the best design approach when it comes to reducing emissions, depending upon the application and design goal.

To demonstrate this, an example is shown in Figure 13 of a spread spectrum clock in the frequency domain in two different designs. The first design has a spectrum that is spread in such a way that the spectral content is wider than the implemented RBW. The second keeps the frequency tight (violating one of our design goals).

Now that we've covered signal spectra, how to measure it, and what spread spectrum modulation is, we'll define components of an average test setup defined in CISPR 25. This test setup can be used to create repeatable measurements from algorithm to algorithm to evaluate different algorithms selected for spreading the clock frequency.

Defining the Average Test Setup

Lastly, to evaluate a design using spread spectrum against CISPR 25 or any recognized standard, we introduce a fixed test setup which consists of:

- A line impedance stabilization network (LISN) or current probe. This device provides a measurement for noise from the device, tuned to the frequency of interest;
- A spectrum analyzer that, when properly configured, provides the measurement capabilities; and
- Copper table or transverse electromagnetic wave (TEM) cell. Both these devices allow for a means of isolation from the environment, with the TEM cell traditionally is used for radiated emissions testing and the copper table for conducted emissions.

This test setup, along with fixed cable lengths, shown in Figure 14, demonstrates a traditional way in how conducted emissions (CE) are measured. It's important to remember, however, to refer to the published standard for an exact setup.

The LISN stabilizes the line impedance, isolating the device from external interference, and provides a measurement port at the specific frequency range of test. Figure 15 demonstrates where the emission from the device can be measured using both a LISN as a measurement device and to stabilize the line impedance.

As demonstrated in these figures, it's not always the power lines that are measured. Depending upon the application, the measurement could include:

- The entirety or portion of a wiring harness
- Positive or return voltage lines
- Control lines

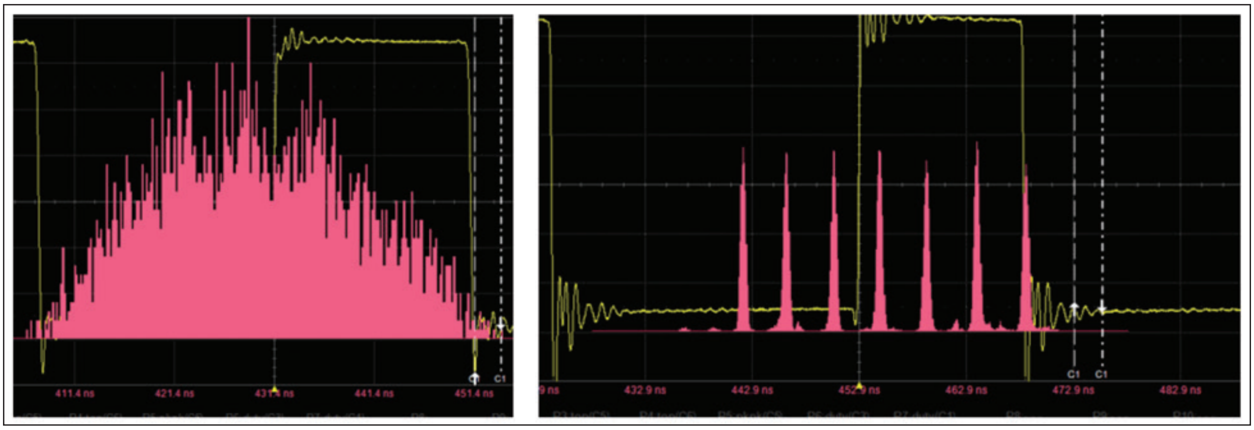


Figure 13: Data showing the difference between a tight modulation scheme (which concentrates energy in a very narrow band close together) and a loose one (which spreads energy out in a very broad band, overlapping each other)

or any combination of these elements. It's important to remember that the ambient noise of the test setup needs to be a significant dB below the intended reference level to receive an accurate measurement. This ambient noise is often taken prior to putting the device under test (DUT) inside of the TEM cell or on the board.

CONCLUSION

When designing or evaluating a design to meet the conducted emissions regulatory requirements found in CISPR 25, it's paramount to have an understanding of the frequency spectra and the measurement of that spectra.

By tracing the conversion process from the continuous time measurement to the frequency domain, we can understand how bandwidth limits and scan times impact the measurement. Next, by understanding how frequency modulation via a spread spectrum algorithm impacts the spectral content, we can overlay these two theories to create design goals. Through these design goals and creating a fixed environment in which to test, proper evaluation can be done on various spread spectrum schemes such that they are able to stay under the various limits imposed by CISPR.

With this understanding, the next time you need to evaluate or design for such a requirement, you have a more complete understanding of how to look at a signal in both the time and frequency domain, and how the spectrum analyzer can be configured to give you the most accurate measurement needed to evaluate your design.

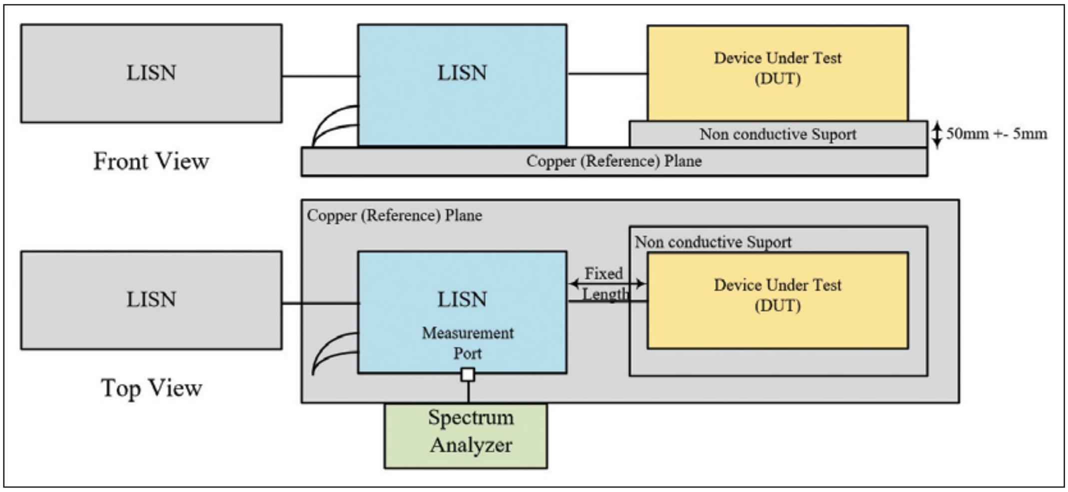


Figure 14: Typical test setup with LISN

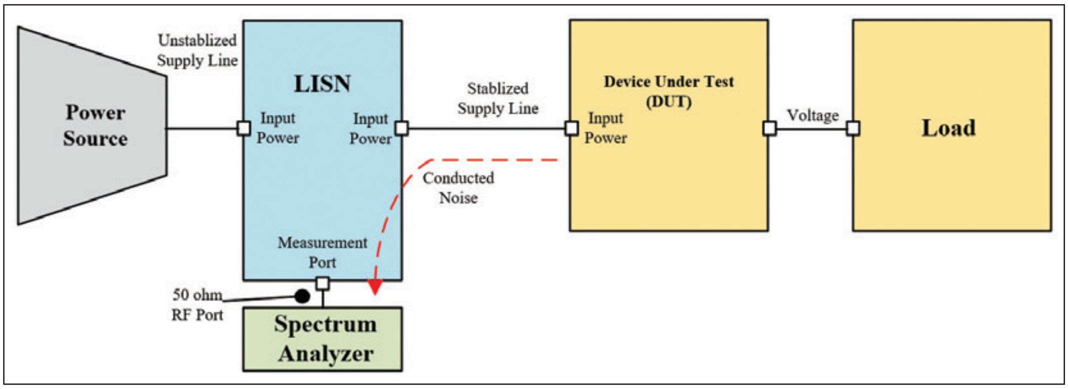


Figure 15: Test setups using a LISN

MIL-STD-464D: A REVIEW OF RECENT CHANGES

A Long-Awaited Update to an Essential Standard for Military Procurement

MIL-STD-464D was released on December 24, 2020. This revision is in keeping with the routine five-year revision cycle applicable to many such standards, and MIL-STD-464 must keep in sync with MIL-HDBK-235, from which the electromagnetic field intensity tables are drawn. In this case, the routine five-year cycle took ten years to complete.

MIL-STD-464 is the U.S. Department of Defense (DoD) top-level E3 requirement set for the procurement of complete or modified systems. In this context, “systems” means an integrated platform of one type or another, such as a ground or air vehicle, a ship or submarine, a spacecraft, or launch vehicle. Note that some systems can be parts of other systems, such as an F-18 fighter aircraft that operates from an aircraft carrier.

The original release of MIL-STD-464 was in 1997. MIL-STD-464A (2002) and MIL-STD-464C (2010) provided minor, evolutionary changes to the original release.[1]

Compared to MIL-STD-464C, the changes in MIL-STD-464D are very minor. This article serves as a laundry list of the substantive changes, including the EME tables, and indications of what values changed in the EME tables, so that the reader may see at a glance where the changes are, rather than checking each table row-by-row and cell-by-cell.

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1. MIL-STD-464C is really MIL-STD-464B, but there was a release cycle error, and MIL-STD-464B was replaced after just a few months. The content didn’t change.

Ken Javor is a Senior Contributor to In Compliance Magazine and has worked in the EMC industry for over 40 years. Javor is an industry representative to the Tri-Service Working Groups that maintain MIL-STD-464 and MIL-STD-461. He can be reached at ken.javor@emccompliance.com.

By Ken Javor

The purpose of this article is to inform and save the reader the time the author spent combing through MIL-STD-464D vs. MIL-STD-464C (referenced as "D" and "C" throughout the rest of this article). Entertaining the reader was not a practical goal.

NEW DEFINITIONS

3.1 All-up-round (AUR)

"The completely assembled munition as intended for delivery to a target or configured to accomplish its intended mission. This term is identical to the term all-up-weapon."

3.2 Bare devices

"Bare electrically initiated devices (EIDs) such as electrical initiators, exploding foil initiators, detonators, etc., in an all-up round that have either one or both pins accessible on an external connector."

3.3 Below deck

Extended to include the pressure hull of a submarine.

3.7 Energetics

"A substance or mixture of substances that, through chemical reaction, is capable of rapidly releasing energy. A few examples of energetics are: liquid and solid propellants such as in rockets and air bags, gun propellants, polymer bonded explosives (PBX) for warheads, pyrotechnics for flares and ignition systems."

3.8 Flight deck

"The upper deck of an aircraft carrier that serves as a runway. The deck of an air-capable ship, amphibious aviation assault ship, or aviation ship used to launch and recover aircraft."

3.12 Helicopter-borne electrostatic discharge (HESD)

"The sudden flow of electric charge between a helicopter or rotary winged aircraft and an object of different electrical potential. A buildup of static electricity can be caused by triboelectric charging or electrostatic induction generated from operating rotary wings."

3.13 High power microwave (HPM)

Deletes the frequency range.

3.18 Maximum no-fire stimulus

MIL-STD-464D	MIL-STD-464C
"The greatest firing stimulus that will not cause initiation or degrade an EID of more than 0.1 % of all electric initiators of a given design at a confidence level of 95%. Stimulus refers to electrical parameters such as current, rate of change of current (di/dt), power, voltage, or energy, which are most critical in defining the no-fire performance of the EID."	*"The greatest firing stimulus which does not cause initiation within five minutes of more than 0.1% of all electric initiators of a given design at a confidence level of 95%. When determining maximum no-fire stimulus for electric initiators with a delay element or with a response time of more than five minutes, the firing stimulus will be applied for the time normally required for actuation."*

3.22 Ordnance (fewer words than "C")

"Explosives, chemicals, pyrotechnics, and similar stores (e.g., bombs, guns, and ammunition, flares, smoke, or napalm)."

3.23 Personnel-borne electrostatic discharge (PESD)

"The sudden flow of electric charge between personnel and an object of different electrical potential. A buildup of static electricity can be caused by triboelectric charging or electrostatic induction generated by the movement of the person's body."

3.27 Spectrum-dependent systems

Adds this statement at the end:

"This includes transmitters, transceivers, and receive-only systems."

3.34 Vertical replenishment (VERTREP)

"The transfer of ordnance and cargo using rotary winged aircraft."

3.35 Weather deck

"The topside of the ship that is exposed to the weather. The weather deck does not include the flight deck, hangar, well deck, man-aloft areas, or the ship's mast."

MAIN BODY REQUIREMENTS

5.1 Margins (MIL-STD-464D)[2]

"Margins shall be established for safety and mission critical subsystems/equipment within the system. Margins shall be no less than 6 dB for safety critical subsystems/equipment, unless otherwise stated in the detailed requirements of this standard. Compliance shall be verified by test, analysis, or a combination thereof."

Compare this with the text in "C," as follows:

"Margins shall be provided based on system operational performance requirements, tolerances in system hardware, and uncertainties involved in verification of system-level design requirements. Safety critical and mission critical system functions shall have a margin of at least 6 dB. EIDs shall have a margin of at least 16.5 dB of maximum no-fire stimulus (MNFS) for safety assurances and 6 dB of MNFS for other applications. Compliance shall be verified by test, analysis, or a combination thereof. Instrumentation installed in system components during testing for margins shall capture the maximum system response and shall not adversely affect the normal response characteristics of the component. When environment simulations below specified levels are used, instrumentation responses may be extrapolated to the full environment for components with linear responses (such as hot bridgewire EIDs). When the response is below instrumentation sensitivity, the instrumentation sensitivity shall be used as the basis for extrapolation. For components with non-linear responses (such as semiconductor bridge EIDs), no extrapolation is permitted."

5.2 Intra-system electromagnetic compatibility (EMC)

MIL-STD-464D	MIL-STD-464C
"The system shall be electromagnetically compatible within itself such that system operational performance requirements are met. Compliance shall be verified by system-level test, analysis, or a combination thereof. This includes permanent, temporary, and portable electronic equipment."	*"The system shall be electromagnetically compatible within itself such that system operational performance requirements are met. Compliance shall be verified by system-level test, analysis, or a combination thereof. For surface ships, MIL-STD-1605(SH) provides test methods used to verify compliance with the requirements of this standard for intra- and inter-system EMC, hull generated intermodulation interference, and electrical bonding."*

5.2.2 Shipboard internal electromagnetic environment (EME)

The very last sentence in "C" section 5.2.2.b after the listing of the individual device and total EIRP is not found in "D." This sentence in "C" that is not in "D" reads:

"Additionally, no device shall be permanently installed within 1 meter of safety or mission critical electronic equipment."

Also, whereas verification in "C" is by test in all cases, in "D," for submarines an analysis consisting of a summation of all individual device EIRP into total radiated power (TRP) is allowed.

(See Tables I – VI, pages 27, 28, and 30)

5.5 Lightning

Has some expanded wording about near strikes and slightly different wording describing Figure 2 and Table VII.

2. Author's note: The significant truncation is due to moving ordnance-related margins to their own separate section. The ordnance margins haven't changed – this just represents a reorganization of the standard.

Frequency Range		Shipboard Flight Decks Electric Field (V/m-rms)		Shipboard Weather Decks Electric Field (V/m-rms)	
(MHz)	(MHz)	Peak	Avg	Peak	Avg
0.01	2	*	*	*	*
2	30	164	164	189/169	189/169
30	150	61	61	61	61
150	225	61	61	61	61
225	400	61	61	61	61
400	700	196	71	445	71
700	790	94	94	94	94
790	1000	491/246	100	744/1307	141/244
1000	2000	212	112	212/112	112
2000	2700	159	159	159	159
2700	3600	4700/2027	595/200	4700/897	595/200
3600	4000	1225/298	200	1859	200
4000	5400	200	200	200	200
5400	5900	361	213	711	235
5900	6000	213	213	235	235
6000	7900	213	213	235	235
7900	8000	200	200	200	200
8000	8400	200	200	200	200
8400	8500	200	200	200	200
8500	11000	913/200	200	913	200
11000	14000	745/744	200	833	200
14000	18000	745/744	200	833	200
18000	50000	200	200	267	200

TABLE I: Maximum external EME for deck operations on Navy ships vs. -464C Table 1. Maximum external EME for deck operations on Navy ships

Frequency Range (MHz)		Main Beam (distances vary with ship class and antenna configuration) Electric Field (V/m –rms)	
		Peak	Avg
0.01	2	*	*
2	30	200	200
30	150	15/10	15/10
150	225	17/10	17/10
225	400	43	43
400	700	2036	268
700	790	20/10	20/10
790	1000	2615/2528	489/485
1000	2000	930	156
2000	2700	21/10	21/10
2700	3600	27460	7500/2620
3600	4000	8553	272
4000	5400	1357/139	198/139
5400	5900	3234	637/267
5900	6000	637/267	637/267
6000	7900	667/400	667/400
7900	8000	667/400	667/400
8000	8400	449/400	449/400
8400	8500	400	400
8500	11000	6900/4173	6900/907
11000	14000	3329	642
14000	18000	3329/3529	642/680
18000	50000	2862	576

‡ The EME levels in the table apply to shipboard operations in the main beam of systems in the 2700 to 3600 MHz frequency range on surface combatants. For all other operations, the unrestricted peak EME level is 12667 V/m and the unrestricted average level is 1533 V/m.

TABLE II: Maximum external EME for ship operations in the main beam of transmitters vs. -464C TABLE 2. External EME for shipboard operations in the main beam of transmitters

-464D values first, -464C values second, where different. Red fill means level has increased. Yellow fill means change is less than 3 dB, either higher or lower, and blue fill means -464D level is lower than for -464C. * means no emitters in that frequency range.

5.7 Subsystems and equipment electromagnetic interference (EMI)

Now includes new wording (in non-italicized in the excerpt that follows):

> *"Individual subsystems and equipment shall meet interference control requirements (such as the conducted emissions, radiated emissions, conducted susceptibility, and radiated susceptibility requirements of MIL-STD-461) so that the overall system complies with all applicable requirements of this standard.* This includes permanent, temporary, and portable electronic equipment. *Compliance shall be verified by tests that are consistent with the individual requirement (such as testing in accordance with MIL-STD-461)."*

Frequency Range (MHz)		Electric Field (V/m-rms)	
		Peak	Avg
0.01	2	1	1
2	30	73	73
30	150	17	17
150	225	4	1
225	400	*	*
400	700	47	6
700	790	1	1
790	1000	7	7
1000	2000	63	63
2000	2700	187	187
2700	3600	23	8
3600	4000	2	2
4000	5400	3	3
5400	5900	164	164
5900	6000	164	164
6000	7900	6	6
7900	8000	3	1
8000	8400	1	1
8400	8500	3	1
8500	11000	140	116
11000	14000	114	114
14000	18000	16	9
18000	50000	23	23

NOTE: *denotes no emitters in that frequency range.

TABLE III: Maximum external EME for space and launch vehicle systems vs. -464C TABLE 3. External EME for space and launch vehicle systems

Frequency Range (MHz)		Electric Field (V/m-rms)	
		Peak	Avg
0.01	2	54/73	54/73
2	30	103	103
30	150	74	74
150	225	41	41
225	400	92	92
400	700	98	98
700	790	58/267	58/267
790	1000	58/284	58/267
1000	2000	232/2452	94/155
2000	2700	638/489	42/155
2700	3600	1148/2450	219
3600	4000	320/489	25/49
4000	5400	645	173/183
5400	5900	5183/6146	129/155
5900	6000	40/549	40/55
6000	7900	3190/4081	292/119
7900	8000	2471/549	296/97
8000	8400	2471/1095	296/110
8400	8500	82/1095	82/110
8500	11000	810/1943	139
11000	14000	3454	102/110
14000	18000	7897/8671	243
18000	50000	2793	48/76

TABLE IV Maximum external EME for ground systems vs. -464c TABLE 4. External EME for ground systems

-464D values first, -464C values second, where different. Red fill means level has increased. Yellow fill means change is less than 3 dB, either higher or lower, and blue fill means -464D level is lower than for -464C. * means no emitters in that frequency range.

5.7.1 Portable electronic devices and carry-on equipment requirements

Newly added in "D," as follows:

> *"Portable electronic devices and carry-on equipment containing electronics which are not permanently installed or integrated into platforms and require airworthiness certification shall meet, as a minimum, the following EMI interface control requirements:*
>
> - *Safety Critical: All platform emissions and susceptibility requirements (such as those defined in MIL-STD-461) that are defined for safety critical equipment.*
> - *Non-Safety Critical: All platform emissions requirements (such as those defined in MIL-STD-461).*
>
> *"If any part of the portable electronic device/carry-on equipment contains radio frequency transmission capability, then transmitter emissions characteristics shall be measured (such as in MIL-STD-461 Test Method CE106), in addition to the applicable requirements stated above. An aircraft EMC evaluation per 5.2 shall also be required to demonstrate platform compatibility of the portable electronic devices/carry-on equipment which have radio frequency transmitting capability.*
>
> *"If any part of the portable electronic device/carry-on equipment contains ordnance or is integrated into an ordnance system, then the HERO requirements stated within this standard shall also be met. Compliance shall be verified by test per the applicable requirements."*

5.7.3 Shipboard DC magnetic field environment. (5.7.2 in "C")

In the "C" revision, this requirement could only be verified by test. In the "D" revision, the ubiquitous phrase, "Compliance shall be verified by test, analysis, or a combination thereof," is used.

5.8.1 Vertical lift and in-flight refueling

Slightly reworded, but the same overall requirement with one significant deletion. The "C" applicability to "any man portable items that are carried internal to the aircraft" has been deleted.

5.8.3 Ordnance subsystems

Rewritten with two brand new sub-paragraphs that break out separately the pre-existing "C" requirement to withstand a 25 kV personnel ESD and adds a separate new requirement to withstand helicopter ESD (300 kV).

Frequency Range (MHz)		Electric Field (V/m – rms)	
		Peak	Avg
0.01	2	200	200
2	30	200	200
30	150	200	200
150	225	200	200
225	400	200	200
400	700	1311	402
700	790	700	183/402
790	1000	700	215/402
1000	2000	6057	232
2000	2700	3351	200
2700	3600	4220	455
3600	4000	3351	657/200
4000	5400	9179	657
5400	5900	9179	657
5900	6000	9179	200
6000	7900	400	200
7900	8000	400	200
8000	8400	7430	266
8400	8500	7430	266
8500	11000	7430	266
11000	14000	7430	558
14000	18000	730	558
18000	50000	1008	200

TABLE V: Maximum external EME for rotary-wing aircraft, excluding shipboard operations vs. -464C Maximum external EME for rotary-wing aircraft, including UAVs, excluding shipboard operations

Frequency Range (MHz)		Electric Field (V/m-rms)	
		Peak	Avg
0.01	2	88	27
2	30	64	64
30	150	67	13
150	225	67	36
225	400	58	3
400	700	2143	159
700	790	554/80	81/80
790	1000	289	105
1000	2000	3363	420
2000	2700	957	209
2700	3600	4220	455
3600	4000	148	11
4000	5400	3551	657
5400	5900	3551	657
5900	6000	148	4
6000	7900	344	14
7900	8000	148	4
8000	8400	187	70
8400	8500	187	70
8500	11000	6299	238
11000	14000	2211	94
14000	18000	1796	655
18000	50000	533	38

TABLE VI: Maximum external EME for fixed-wing aircraft, excluding shipboard operations vs. -464C TABLE 6. External EME for fixed wing aircraft, including UAVs, excluding shipboard operations

-464D values first, -464C values second, where different. Red fill means level has increased. Yellow fill means change is less than 3 dB, either higher or lower, and blue fill means -464D level is lower than for -464C. * means no emitters in that frequency range.

5.8.4 Electrical and electronic subsystems

Rewritten to refer to MIL-STD-461G (CS118) for test, whereas previously they had to point elsewhere.

5.9.3 Hazards of electromagnetic radiation to ordnance (HERO)

Rewritten to include ordnance safety margins that were struck from general margin paragraph 5.1.

5.14.2 Platform radiated emissions

Renamed from the same paragraph in "C" labeled *5.14.2 Inter-system EMC*. The requirement has both greater generality and is more specific about what parameters need to be controlled. New sub-paragraph in "D."

6.2 Acquisition requirements

Acquisition documents should specify the following: a. Title, number, and date of this standard.

6.3 DIDs

Not updated.

6.5 Key Words

Adds two new terms, electrostatic and HESD.

6.6 International standardization agreement implementation.

Rewritten slightly in "D" from the previous similar section 6.5 in "C."

6.7 Acronyms

Replaces "EMRADHAZ" with "RADHAZ." Also, PESD and HESD are added.

6.8 Technical points of contact

Air Force and Army points-of-contact have been updated.

Frequency Range		Field Intensity (V/m – rms)			
		Unrestricted*		Restricted **	
(MHz)	(MHz)	Peak	Avg	Peak	Avg
0.01	2	200	200	80	80
2	30	200	200	100	100
30	150	200	200	80	80
150	225	200	200	70	70
225	400	200	200	100	100
400	700	2200	410	450	100
700	790	700	190	270	270
790	1000	2600	490	1400	270
1000	2000	6100	420	2500	160
2000	2700	6000	500	490	160
2700	3600	27460	5350/2620	2500	220
3600	4000	8600	280	1900	200
4000	5400	9200	660	650	200
5400	5900	9200	660	6200	240
5900	6000	9200	640/270	550	240
6000	7900	3190/4100	670/400	3190/4100	240
7900	8000	2500/550	670/400	550	240/200
8000	8400	7500	450/400	1100	200
8400	8500	7500	400	1100	200
8500	11000	7500	3450/910	2000	300
11000	14000	7500	650/680	3500	220
14000	18000	7900/8700	650/680	7900/8700	250
18000	50000	2900	580	2800	200

NOTES:

*It must be noted that on certain naval platforms, there are radar systems (and unique modes of operation) that may produce fields in excess of those in Table IX, and MIL-HDBK-235 must be consulted to identify specific EME test requirements.

** In some of the frequency ranges for the "Restricted Average" column, limiting the exposure of personnel through time averaging will be required to meet the requirements of 5.9.1 for personnel safety.

TABLE IX: Maximum external EME levels for ordnance vs. -464C TABLE 9. Maximum external EME levels for ordnance.

-464D values first, -464C values second, where different. Red fill means level has increased. Yellow fill means change is less than 3 dB, either higher or lower, and blue fill means -464D level is lower than for -464C. * means no emitters in that frequency range.

APPENDICES AND GUIDANCES

A.1.1 Scope

Includes extra language emphasizing that appendix is guidance only, not mandatory.

A.2.1.1 Specifications, standards, and handbooks

Slightly different wording. Also, the following additions, changes, and deletions:

- MIL-STD-1576, Electroexplosive Subsystem Safety Requirements and Test Methods for Space Systems—removed from applicable documents
- MIL-STD-3023 HEMP Protection for Military Aircraft—added
- MIL-STD-4023 HEMP Protection for Maritime Assets—added
- MIL-HDBK-83578 Criteria for Explosive Systems and Devices Used on Space Vehicles—deleted

A.2.1.2 Other Government documents, drawings, and publications

- Army, ATPD-2407 Electromagnetic Environmental Effects (E3) for U.S. Army Tank and Automotive Vehicle Systems Tailored from MIL-STD-464C—added
- TOP 01-2-511A US Army Test and Evaluation Command Test Operations Procedure—added

A.2.2 Non-Government Publications

- Institute of Electrical and Electronics (IEEE) Transactions on Electromagnetic Compatibility
- DOI:10.1109/TEMC.2016.2575842 Effect of Human Activities and Environmental Conditions on Electrostatic Charging—added
- Franklin Applied Physics
- F-C2560 RF Evaluation of the Single Bridgewire Apollo Standard Initiator—deleted

A.3 Acronyms

- AMITS air management information tracking system—deleted
- EMRADHAZ—deleted
- HESD helicopter-borne electrostatic discharge—added
- PESD personnel-borne electrostatic discharge—added
- RADHAZ Radiation hazards—added

A.4.1 Requirement Guidance

Adds Army ATPD-2407 and TOP 01-2-511A is EMC guidance and test procedures.

A.4.1.e Requirement Guidance

Includes additional guidance and a slightly different approach than "C." Margin Requirement Guidance A.5.1 adds the non-italicized statement in the following excerpt:

> *"Margins need to be viewed from the proper perspective. The use of margins simply recognizes that there is variability in manufacturing and that requirement verification has uncertainties. The margin ensures that every produced system will meet requirements, not just the particular one undergoing a selected verification technique. Smaller margins are appropriate for situations where production processes are under tighter controls or more accurate and thorough verification techniques are used. Smaller margins are also appropriate if many production systems undergo the same verification process, since the production variability issue is being addressed. Margins are not an increase in the basic defined levels for the various electromagnetic environments. The most common technique is to verify that electromagnetic and electrical stresses induced internal to the system by external environments are below equipment strength by at least the margin.* This approach is similar to the test methodology described in A.4.1 (e). *While margins can sometimes be demonstrated by performing verification at a level in excess of the defined requirement, the intent of the margin is not to increase the requirement."*

This paragraph is deleted from this section in "D" (look for it in the EID section):

> *"MNFS values for EIDs are normally specified by manufacturers in terms such as DC currents or energy. Margins are often demonstrated by observing an effect during the application of an electromagnetic environment that is the same effect observed when applying a stimulus level in the form under which the MNFS is defined. For example, the temperature rise of a bridgewire can be monitored in the presence of an EME*

relative to the temperature rise produced by a DC current level that is 16.5 dB below MNFS. The space community has elected to use MNFS levels determined using RF rather than DC. This approach is based on Franklin Institute studies, such as report F-C2560. Outside of the space community, the use of DC levels has provided successful results."

A.5.2 Intra-system EMC

Under *Requirements Rationale***,** the final sentence in "C":

> *"To ensure EMC is achieved in Navy ships, a MIL-STD-1605(SH) survey should be performed."*

is replaced by a more descriptive version in "D":

> *"For surface ships, MIL-STD-1605(SH) provides test methods used to verify compliance with the requirements of this standard for intra- and inter-system EMC, hull generated intermodulation interference, and electrical bonding."*

A.5.2 Verification Guidance

The following and final line item is modified in "D" to read:

> *"For portable electronic devices and carry-on equipment, EMI requirements are defined in 5.7.1."*

In "C," line item h reads:

> *"TABLE A- 1 identifies what kind of EMI/EMC testing is required when new, modified, or carry-on equipment will be used on military aircraft."*

Table A-1 Type of EMI/EMC testing doesn't exist in "D."

A.5.3 Requirement Guidance

These words added to the very end of this section:

> *"A platform design, while descriptively fitting the title of an external EME table (e.g., Fixed Wing or Rotary Wing), may not coincide with the platform's operational EME definition. Strict attention must be paid to the assumptions used in deriving the tables to ensure appropriate EMC compliance."*

A.5.4 Requirement Guidance (HPM)

Eliminates Tables A-4 – A-10 from "C" and also calculation of some example problems using these tables.

A.5.4 Requirement Rationale (HPM)

Eliminates some wording questioning the effectiveness of HPM.

A.5.4 Verification Guidance (HPM)

Eliminates reference to these deleted examples in "D."

A.5.6 Requirement Guidance (EMP)

Contains some extra description of HEMP composite environment. It also adds descriptions of EMP-related military standards for dealing with EMP, including effects on spacecraft.

A.5.6 Requirement Lessons Learned

Has this sentence in common with "C":

> *"Hardening against ground-burst nuclear radiation environments is often not cost effective because a burst near enough to produce a radiation and electromagnetic threat is also close enough for the blast to disable the facility."*

But "D" adds this last sentence not in "C":

> *"Buried facilities such as ICBM launch sites are an exception."*

A.5.6 Verification Rationale (EMP)

"D" replaces this "C" paragraph:

> *"For many systems, the cost of EMP verification is a major driver. Therefore, the procuring activity should decide what level of verification is consistent with the risk that they are willing to take."*

with this paragraph:

> *"High-altitude EMP protection standards have been developed for fixed ground-based facilities, transportable ground-based systems, aircraft and ships. Each of these standards contains detailed verification testing protocols and pass/fail criteria. Use of these standards is mandatory for DoD military system procurements that have a HEMP requirement."*

Note the emphasis on the cost of EMP design has been replaced with wording more conducive to getting EMP designs installed.

In the same section, this new "D" wording:

> *"MIL-STD-3023 and MIL-STD-4023 for HEMP protection of military aircraft and ships, respectively provide a similar verification test approach except that these standards require illuminating the aircraft and ships with a simulated plane wave HEMP threat environment and measuring the induced stresses at each MCS equipment interface. Each MCS must be tested to MIL-STD-461 CS116 to establish its immunity before being installed into the platform. A user selectable margin is then applied to the measured current stress which is then pulse current injected (PCI) at the same interface used in the MIL-STD-461 CS116 testing. This enables direct stress to immunity comparisons at common interfaces for each mission critical equipment throughout the system. Monitoring for upset and damage is also performed at this time."*

has been appended to this existing "C" wording:

> *"MIL-STD-188-125-1 and MIL-STD-188-125-2 contain verification test methods for demonstrating that C^4I fixed ground-based facilities and transportable ground-based systems meet HEMP requirements. The test methods describe coupling of threat-relatable transients using pulse current injection to penetrating conductors at injection points outside of the facility shield."*

A.5.7 Requirement Guidance (Subsystem & Equipment EMI)

Eliminates wording about DO-160 section 22 now that CS117 is available.

A.5.7.1 Portable Electronic Devices and Carry-On Equipment Requirements

All new appendix material. Basically refers to A.5.2. Intra-system EMC.

A.5.8.1 Vertical lift and in-flight refueling

Slightly rewritten, no changes.

A.5.8.3 Ordnance Subsystems

Greatly expanded and also includes the following new sections:

- A.5.8.3.1 Personnel-borne ESD (PESD) for ordnance and ordnance systems

- A.5.8.3.2 Helicopter-borne ESD (HESD) for ordnance and ordnance systems

A.5.9.3 Requirement Rationale (Ordnance RADHAZ (HERO)).

This section is rewritten with substantive changes.

A.5.9.3 Requirement Guidance (Ordnance RADHAZ (HERO))

This section is rewritten with substantive changes. MIL-STD-464C was:

> *"OD 30393 provides design principles and practices for controlling electromagnetic hazards to ordnance. MIL-STD-1576 and MIL-HDBK-83578 (USAF) provide guidance on the use of ordnance devices in space and launch vehicles. For space applications using ordnance devices, an analysis of margins based on the RF threshold determination of the MNFS should be performed."*

The last sentence refers to measuring the rf TOS of bridgewires, and that has been completely debunked. This section now reads:

> *"NASA document TP2361 provides design guidelines for space and launch vehicle charging issues. Subsystems and equipment installed aboard space systems should be able to meet operational performance requirements during and/or after being subjected to representative discharges simulating those due to spacecraft charging."*

A.5.14.2 Requirement Rationale (Platform Radiated Emission)

Rewritten with added information.

A.5.15 Requirement Guidance (EM Spectrum Compatibility)

Completely rewritten.

A.5.15 Verification Rationale (EM Spectrum Compatibility)

Completely rewritten.

A.5.15 Verification Guidance (EM Spectrum Compatibility)

Added information.

HOW GROUNDS AFFECT THE PEAK VOLTAGE DUE TO LIGHTNING

Why the Most Common Characterization of a Ground Rod May Not Work for Lightning

Al Martin holds a BEE degree from Cornell University and a Ph.D. from UCLA. He is the author or co-author of over 35 papers on EMC and telecommunications and is a Life Senior Member of the IEEE and the IEEE SA. Martin is also interested in particle physics and is presently part of a voluntary computing network serving the European Center for Nuclear Research. He can be reached at amartin_36@yahoo.com.

By Albert R. Martin

In 1997, an experiment at the Camp Blanding center for lightning testing [1] challenged the predominant view that ground rods are essentially resistive. What that experiment found was that the waveshapes of lightning currents in a building grounding system and those entering the electrical circuits of the building were considerably different. That was at odds with IEC 61312-1:1995 [2] assertions that they should be the same. The conclusion was that, for lightning, the ground rod had an impedance with a reactive component in addition to the resistive one.

So how do we take into account the impedance effects for lightning? Well, it turns out not to be so simple. Professor Leonid Grcev, who with his students has conducted extensive studies of grounds, has found that a simple modeling of a ground rod as an R-L-C circuit doesn't give correct results, due to surge propagation effects which cause a deviation from the low frequency behavior during the fast-transient period. So the challenge is to determine what this deviation is.

Considering normal grounds (those not chemically treated or otherwise enhanced), Grcev has shown that they can be characterized in terms of effective length and impulse coefficient (IC) [3]. The IC is the ratio of peak voltage across an actual ground rod to the peak voltage across a purely resistive ground rod in response to a surge. It shows how the impedance of the ground rod affects the expected peak voltage due to a surge relative to what it would have been if the ground rod were purely resistive.

EFFECTIVE LENGTH

The first thing to consider is the ground rod effective length l_{eff}, which is the maximum length of the ground electrode for which the impulse coefficient is equal to one. l_{eff} will be used later in the discussion of the IC (which is what we really want).

To calculate l_{eff}, Grcev [3] has developed the relation:

$$l_{eff} = \frac{1-\beta}{\alpha} \quad (1)$$

where:

$$\alpha = 0.025 + exp[-0.82(\rho \cdot T_1)^{0.257}] \quad (2)$$

$$\beta = 0.17 + exp[-0.22(\rho \cdot T_1)^{0.555}] \quad (3)$$

ρ = soil resistivity in *ohm-m* and T_1 is the zero-to-peak rise time of the lightning current pulse. MIL-HDBK-419 Table 2.3 [6] shows a range for average soil resistivity of 1 to 500 ohm-m. CIGRE TB549 Table 3.5 [7] shows a range of front durations of 1.1 µsec for the average subsequent stroke to 18 µsec for the maximum first stroke. Considering those values, the ρT_1 product could reasonably range from 1 to over 1000 ohm-m-µsec. We can use those values in equations (2) and (3) to make a plot of l_{eff} vs. ρT_1, as shown in Figure 1. Both slower rise-time and higher soil resistivity lead to a longer effective ground-rod length.

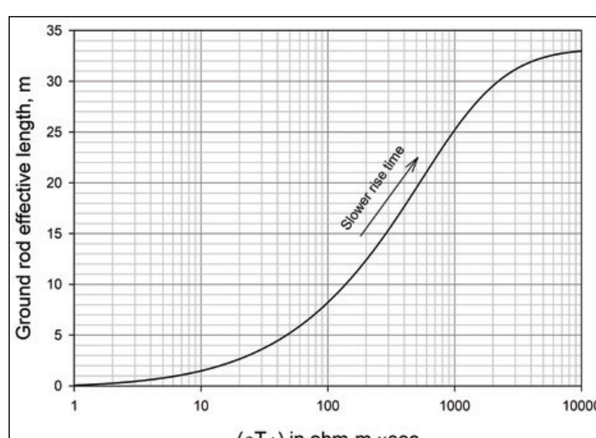

Figure 1: This figure shows the variation in the effective length of a ground rod with soil resistivity and the zero-to-peak time of the surge.

IMPULSE COEFFICIENT

If the length s of the ground rod is less than l_{eff} (see Figure 1), the ground rod is primarily resistive, with some capacitive effect. If the length of the ground rod is greater than l_{eff}, the ground rod will have inductive effects. So which effect do we have, and what is the consequence of that effect? Well, that's what the IC determines. Grcev [3] has proposed the relation:

$$A = \alpha s + \beta \tag{4}$$

where A = Z/R is the impulse coefficient, Z is the effective impedance, R is the ground rod resistance, α is calculated from equation (2), and β is calculated from equation (3).

For A > 1, the ground rod has an effective series inductance in addition to its resistance. In this case, the peak voltage will be A times bigger than it would have been if the ground rod were purely resistive.

For A < 1, the ground rod has an effective parallel capacitance in addition to its resistance. In this case, the peak voltage will be A times lower than it would have been if the ground rod were purely resistive.

From equation (4) the effect of the ground rod reactance can be calculated. As an illustration, take the four cases of ρT_1, = 100, 300, 1000, and 10,000, and use equation (4) to plot the impulse coefficient A vs. length of the rod. Ground rods with a low ρT_1 product have a high impulse coefficient, whereas ground rods with a high ρT_1 product have a low impulse coefficient, as shown in Figure 2.

Figure 3 is a replot of Figure 2 for ground rods of a length normally used (≤ 10 m).

For ground rods ≤ 10 m, the low value of the impulse coefficient means that the peak voltage across the ground rod will be less than would be calculated for a purely resistive ground rod. For example, for a common 2 m rod, the ratio of peak voltage to the peak voltage across a purely resistive ground rod is in the range of 0.2 to 0.4, depending on the ρT_1 product. The voltage across the ground rod as a surge decays is determined primarily by the resistance of the ground rod. So as the surge decays, the effect of the ground rod reactance dies away (remember that the impulse coefficient is relevant only during the rise-time period).

CURRENT FLOWING IN THE GROUND ROD

The peak voltage developed across the ground rod is given by:

$$V_{peak} = ZI_{rod} \tag{5}$$

where I_{rod} is the peak current captured by the ground rod, and Z is the ground rod impedance.

To calculate I_{rod} we need to calculate the fraction of the lightning current I_{max} captured by the ground rod. IEEE Std 142 [5] shows that 99% of the current flowing in the ground rod is captured in a volume having a radius of twice a ground rod length, s. Figure 4 illustrates this situation, where d is the distance from the lightning strike point to the edge

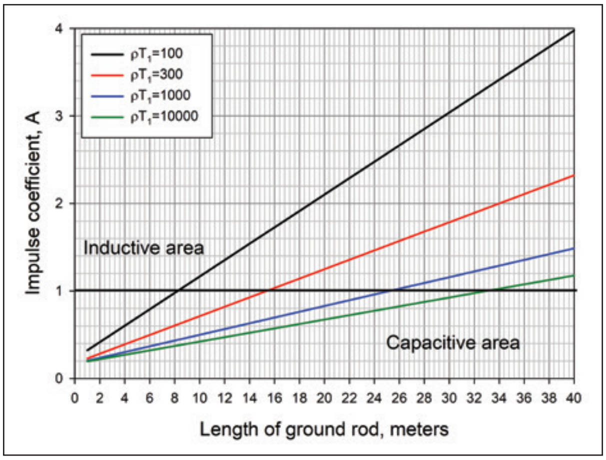

Figure 2: Impulse coefficient (ratio of peak voltage to the peak voltage across a purely resistive ground rod) versus length of ground rod

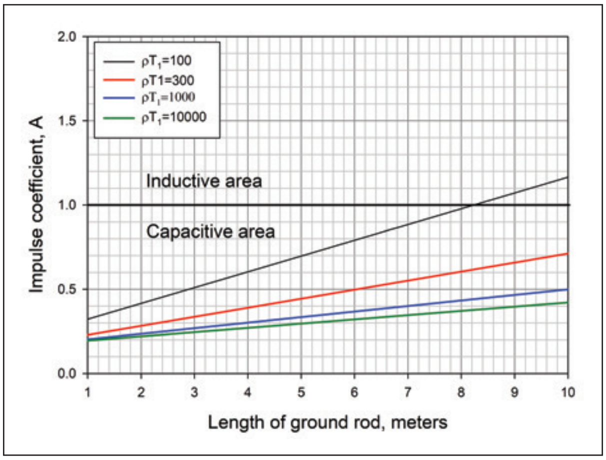

Figure 3: Impulse coefficient for ground rods ≤ 10 m long

of a cylinder representing the ground rod outer effective extent.

The angle θ subtended by the ground rod is given by:

$$\theta = 2\arcsin\left(\frac{2s}{d}\right) \tag{6}$$

Note that the arcsin is not defined for arguments greater than 1, so there are two cases for equation (6): Case 1 where d ranges from $2s$ to infinity, and case 2 where d ranges from $2s$ to 0.

For case 1, if the arcsin is in degrees, then the fraction f_1 of the lightning current I_{max} captured by the ground rod is:

$$f_1 = \frac{2\arcsin\left(\frac{2s}{d}\right)}{180} = \frac{\arcsin\left(\frac{2s}{d}\right)}{90} \tag{7}$$

For case 2, if the fraction f_2 of the lightning current I_{max} captured by the ground rod is:

$$f_2 = \frac{\arcsin\left(1-\frac{d}{2s}\right)}{90} \tag{8}$$

Combining equations (7) and (8), $I_{rod} = I_{max}\,(f_1 + f_2)$, which is:

$$I_{rod} = I_{max}\left\{\left[\frac{\arcsin\left(\frac{2s}{d}\right)}{90}\right] + \left[\frac{\arcsin\left(1-\frac{d}{2s}\right)}{90}\right]\right\} \tag{9}$$

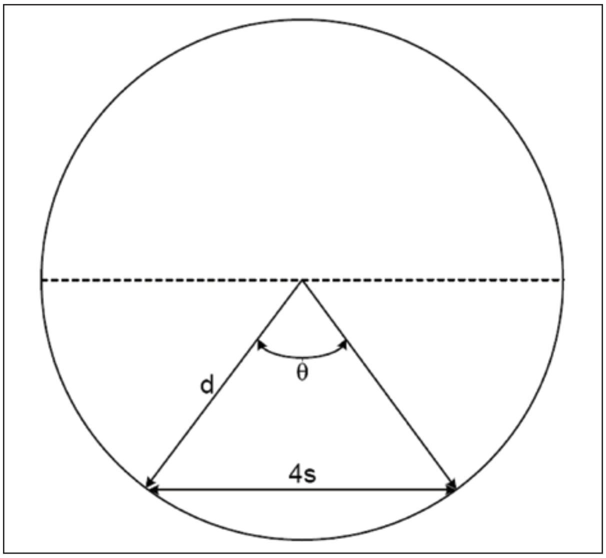

Figure 4: The effective capture area of the ground rod

Remember that in calculating I_{rod}, the first term in equation (9) is only valid for d greater than $2s$, and the second term is only valid for d less than $2s$.

PEAK VOLTAGE

The peak voltage is calculated from equation (5). The effective impedance Z of the ground rod to be used in equation (5) can be calculated from Dwight's [4] equation multiplied by A:

$$Z = \frac{A\rho}{2\pi s}\left[\ln\left(\frac{4s}{a}\right) - 1\right] \tag{10}$$

where a is the radius of the ground rod.

Substituting equations (9) and (10) in equation (5):

$$V_{peak} = I_{max}\left\{\left[\frac{\arcsin\left(\frac{2s}{d}\right)}{90}\right] + \left[\frac{\arcsin\left(1-\frac{d}{2s}\right)}{90}\right]\right\}\left\{\frac{A\rho}{2\pi s}\left[\ln\left(\frac{4s}{a}\right) - 1\right]\right\}$$

$$V_{peak} = \frac{A\rho I_{max}\left[\arcsin\left(\frac{2s}{d}\right)+\arcsin\left(1-\frac{d}{2s}\right)\right]}{180\pi s}\left[\ln\left(\frac{4s}{a}\right) - 1\right] \tag{11}$$

As an example of the calculation of V_{peak}, consider a 12 kA 4.5/77 subsequent surge from TB549 [7] impinging on a 10 m rod 5/8 inches in diameter in the soil of 50 ohm-cm, 200 ohm-cm 600 ohm-cm, and 3000 ohm-cm.

For these cases, Figure 5 shows how V_{peak} changes due to a decrease in ground-rod current capture with increasing distance.

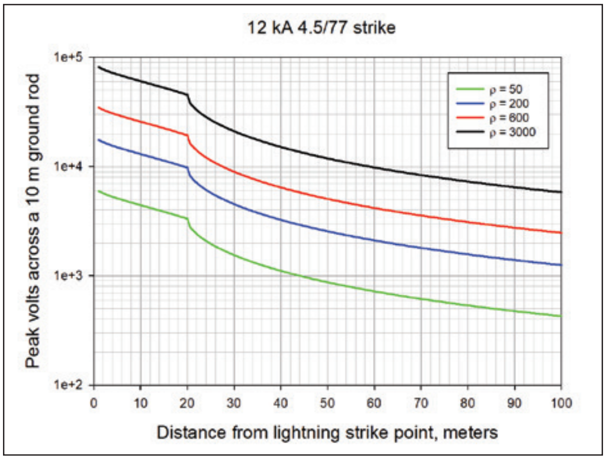

Figure 5: Example of the peak voltage across a 2 m ground rod due to a 12 kA 4.5/77 strike

APPLICABILITY OF THE PEAK VOLTAGE CALCULATION

Now a word about the applicability of the foregoing analysis. In the region near the lightning strike point, the ground resistivity ρ is highly variable. In particular, soil breakdown can happen when the electric field overcomes the soil ionization gradient [8]. Soil ionization occurs when the electric fields at the ground electrode surface become greater than the ionization threshold of approximately 300 kV/m [9]. In this case, in the region surrounding the current striking point, local transverse discharges start from the lightning strike point and stop at the points where the electric field drops below the critical breakdown strength. An illustration of this point is shown in Figure 6.

The literature on lightning shows that the streaks in Figure 6 are places where the ground is ionized. A circle of radius r_0 can be put around this area. The size of r_0 is determined by both the magnitude of the lightning current and ρ. In Figure 6, r_0 appears to be about 6 m, but that may or may not be typical. In any case, to avoid the area where ρ is highly variable, *d* should generally exceed $2r_0$.

With the foregoing discussions in mind, different lightning waveforms, different ρ, and different ground rod lengths will result in different peak voltages from those shown in Figure 5.

SUMMARY

The usual assumption that ground rods are purely resistive is actually not what is observed in the case of lightning. Particularly for the relatively short ground rods commonly used, during the rise-time period the ground rods look like an impedance with a significant capacitive component. The result is that for these commonly used ground rods, the peak voltage due to a lightning strike is generally significantly lower than would be the case for a purely resistive ground rod. Whether the peak voltage is higher or lower than for a purely resistive ground rod depends on a number of variables, including the surge waveform, the ground resistivity, the length of the ground rod, and the distance the observer is from the lightning strike point. The peak voltage across the ground rod can be calculated, based on estimates of these variables.

Figure 6: Extent of ionization from a lightning strike to the flag marking the hole

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EM WAVES, VOLTAGE, AND CURRENT WAVES

By Bogdan Adamczyk

This article presents a concept of a wave together with the wave equations and their solutions. The time-varying EM fields and their propagation in both the time and frequency domains is discussed first. Subsequently, the equations for the voltages and currents on the transmission line are obtained. It is shown that these equations and that their solutions represent voltage and current waves propagating along the line.

Dr. Bogdan Adamczyk is professor and director of the EMC Center at Grand Valley State University (http://www.gvsu.edu/emccenter) where he regularly teaches EMC certificate courses for industry. He is an iNARTE certified EMC Master Design Engineer. Prof. Adamczyk is the author of the textbook "Foundations of Electromagnetic Compatibility with Practical Applications" (Wiley, 2017) and the upcoming textbook "Principles of Electromagnetic Compatibility with Laboratory Exercises" (Wiley 2022). He can be reached at adamczyb@gvsu.edu.

1. CONCEPT OF A WAVE

Consider a function of time t and space z, , with its argument given by

$$f(z,t) = f\left(t - \frac{z}{v}\right) \tag{1.1}$$

Then,

$$f(z+\Delta z, t+\Delta t) = f\left(t + \Delta t - \frac{z+\Delta z}{v}\right) \tag{1.2}$$

Eq. (1.2) is valid for any Δz and any Δt. Thus, we could choose any relationship between the two deltas, and the new equation would still be valid. Let's choose this relationship to be

$$\Delta z = v\Delta t \tag{1.3}$$

Then, Eq. (1.2) becomes

$$f(z+\Delta z, t+\Delta t) = f\left(t + \Delta t - \frac{z+v\Delta t}{v}\right) = f\left(t + \Delta t - \frac{z}{v} - \Delta t\right) = f\left(t - \frac{z}{v}\right) \tag{1.4a}$$

or

$$f(z+\Delta z, t+\Delta t) = f\left(t - \frac{z}{v}\right) \tag{1.4b}$$

Therefore, after a time Δt, the function f retains the same value at a point that is $\Delta z = v\Delta t$ away from the previous position in space (defined by z), as shown in Figure 1.

This means that any function of the form represents a wave traveling in the positive z direction with a velocity

$$v = \frac{\Delta z}{\Delta t} \tag{1.5}$$

Similarly, it can be shown that any function of the form $f(t - z/v)$ represents a wave traveling in the negative z direction as the time advances.

2. UNIFORM PLANE EM WAVE IN TIME DOMAIN

The time variations of the magnetic (***H***) and electric (***E***) fields give rise to the space variations of the electric and magnetic fields, respectively. This interdependence of the space and time variations gives rise to the electromagnetic wave propagation.

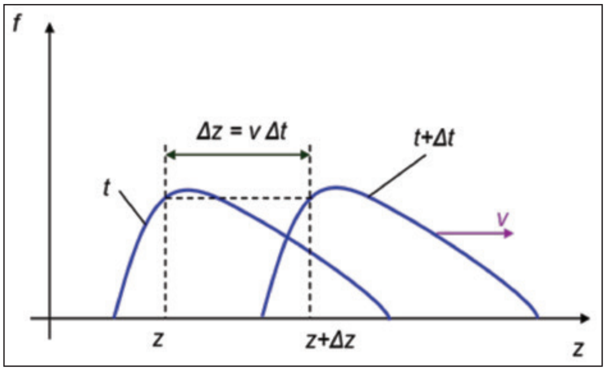

Figure 1: Wave propagating in the positive z direction with a velocity v

The two fields are related by Maxwell's equations (in source-free medium)

$$\nabla \times \boldsymbol{E} = -\mu \frac{\partial \boldsymbol{H}}{\partial t} \tag{2.1a}$$

$$\nabla \times \boldsymbol{H} = \sigma \boldsymbol{E} + \varepsilon \frac{\partial \boldsymbol{E}}{\partial t} \tag{2.1b}$$

where μ, σ, and ε are the permeability, conductivity, and permeability of a medium, respectively.

In general, the electric and magnetic fields have three nonzero components, each of them being a function of all three coordinates and time. That is,

$$\boldsymbol{E} = [E_x(x,y,z,t), E_y(x,y,z,t), E_z(x,y,z,t)] \tag{2.2a}$$

$$\boldsymbol{H} = [H_x(x,y,z,t), H_y(x,y,z,t), H_z(x,y,z,t)] \tag{2.2b}$$

We will focus on a simple and very useful type of wave: *uniform plane wave*. Uniform plane waves not only serve as a building block in the study of electromagnetic waves but also support the study of wave propagation on transmission lines as we will show.

Under the uniformity in the plane assumption, if the ***E*** field points in the $+x$ direction (usual designation) then the Maxwell's equations show that the ***H*** field is pointing in the $+y$ direction, and

$$\boldsymbol{E} = [E_x(z,t), 0, 0] \tag{2.3a}$$

$$\boldsymbol{H} = [0, H_y(z,t), 0] \tag{2.3b}$$

This is shown in Figure 2.

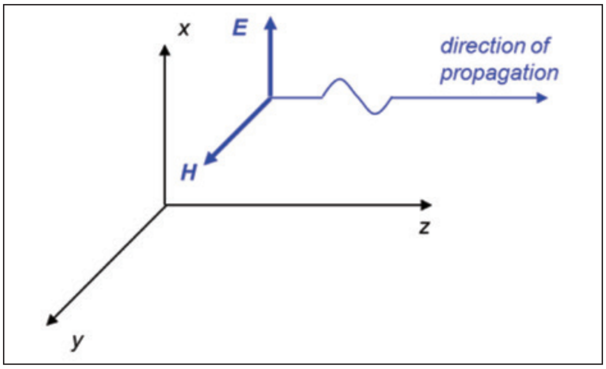

Figure 2: Uniform plane EM wave

The fields propagate as waves in the positive $+z$ direction. Under the uniformity in the plane assumption, Equations (2.1) for a lossless medium (σ = 0) become

$$\frac{\partial^2 E_x(z,t)}{\partial z^2} = \mu\varepsilon \frac{\partial^2 E_x(z,t)}{\partial t^2} \tag{2.4a}$$

$$\frac{\partial^2 H_y(z,t)}{\partial z^2} = \mu\varepsilon \frac{\partial^2 H_y(z,t)}{\partial t^2} \tag{2.4b}$$

and their general solution, in a lossless medium, is [1],

$$E_x(z,t) = Af\left(t - \frac{z}{v}\right) + Bg\left(t + \frac{z}{v}\right) \tag{2.5a}$$

$$H_y(z,t) = \frac{A}{\eta} f\left(t - \frac{z}{v}\right) - \frac{B}{\eta} g\left(t + \frac{z}{v}\right) \tag{2.5b}$$

where

$$\eta = \sqrt{\frac{\mu}{\varepsilon}} \quad (\Omega) \tag{2.6}$$

is the *intrinsic impedance* of a (lossless) medium, and A and B are constants.

Based on the discussion in Section 1, we recognize the functions f and g, as waves propagating in $+z$ and $-z$ directions, respectively, with a velocity of propagation equal to

$$v = \frac{\Delta z}{\Delta t} = \frac{1}{\sqrt{\mu\varepsilon}} \tag{2.7}$$

3. UNIFORM PLANE EM WAVE IN FREQUENCY DOMAIN

In the previous section, we described the wave equations in a lossless medium for arbitrary time variations. When the time variations are sinusoidal, the wave equations in any (simple) medium become [1]:

$$\frac{d^2 \hat{E}_x}{dz^2} = \hat{\gamma}^2 \hat{E}_x(z) \tag{3.1a}$$

$$\frac{d^2 \hat{H}_y}{dz^2} = \hat{\gamma}^2 \hat{H}_y(z) \tag{3.1b}$$

where

$$\hat{\gamma} = \sqrt{j\omega\mu(\sigma + j\omega\varepsilon)} \tag{3.2}$$

is the *propagation constant* of the medium. The general solution of Equations (3.1) is

$$\hat{E}_x = E_m^+ e^{-\hat{\gamma} z} + E_m^- e^{\hat{\gamma} z} \tag{3.3a}$$

$$\hat{H}_y = \frac{E_m^+}{\hat{\eta}} e^{-\hat{\gamma} z} - \frac{E_m^-}{\hat{\eta}} e^{\hat{\gamma} z} \tag{3.3b}$$

where

$$\hat{\eta} = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\varepsilon}} = \frac{j\omega\mu}{\hat{\gamma}} = \eta\angle\theta_\eta \quad [\Omega] \tag{3.4}$$

is the *intrinsic impedance* of the medium. The propagation constant is often expressed in terms of its real and imaginary parts as

$$\hat{\gamma} = \alpha + j\beta \tag{3.5}$$

where α is the attenuation constant and β = w/*v* is the phase constant. The complex intrinsic impedance is often expressed in an exponential form as

$$\hat{\eta} = \eta e^{j\theta_\eta} \tag{3.6}$$

Then the solution in Equations (3.3) can be written as

$$\hat{E}_x = \hat{E}_m^+ e^{-\alpha z} e^{-j\beta z} + \hat{E}_m^- e^{\alpha z} e^{j\beta z} \tag{3.7a}$$

$$\hat{H}_y = \frac{\hat{E}_m^+}{\eta} e^{-\alpha z} e^{-j\beta z} e^{-j\theta_\eta} - \frac{\hat{E}_m^-}{\eta} e^{\alpha z} e^{j\beta z} e^{-j\theta_\eta} \tag{3.7b}$$

Often, the undetermined complex constants can be expressed as

$$\hat{E}_m^+ = E_m^+ \angle 0 = E_m^+ \tag{3.8a}$$

$$\hat{E}_m^- = E_m^- \angle 0 = E_m^- \tag{3.8b}$$

Then, the solutions in Equations (3.7) become

$$\hat{E}_x = E_m^+ e^{-\alpha z} e^{-j\beta z} + E_m^- e^{\alpha z} e^{j\beta z} \tag{3.9a}$$

$$\hat{H}_y = \frac{E_m^+}{\eta} e^{-\alpha z} e^{-j\beta z} - \frac{E_m^-}{\eta} e^{\alpha z} e^{j\beta z} \tag{3.9b}$$

The corresponding time-domain solutions, in a *lossless medium*, are [1]:

$$E_x = E_m^+ \cos(\omega t - \beta z) + E_m^- \cos(\omega t + \beta z) \tag{3.10a}$$

$$H_y = \frac{E_m^+}{\eta} \cos(\omega t - \beta z) - \frac{E_m^-}{\eta} \cos(\omega t + \beta z) \tag{3.10b}$$

Note that

$$\cos(\omega t - \beta z) = \cos\omega\left(t - \frac{\beta}{\omega} z\right) = \cos\omega\left(t - \frac{z}{v}\right) = f\left(t - \frac{z}{v}\right) \tag{3.11a}$$

$$\cos(\omega t + \beta z) = \cos\omega\left(t + \frac{\beta}{\omega} z\right) = \cos\omega\left(t + \frac{z}{v}\right) = g\left(t + \frac{z}{v}\right) \tag{3.11b}$$

Thus, the Equations (3.10) represents sinusoidal traveling wave in the +*z*, and -*z* direction, respectively!

Figure 3 shows a forward propagating EM wave in a lossless medium.

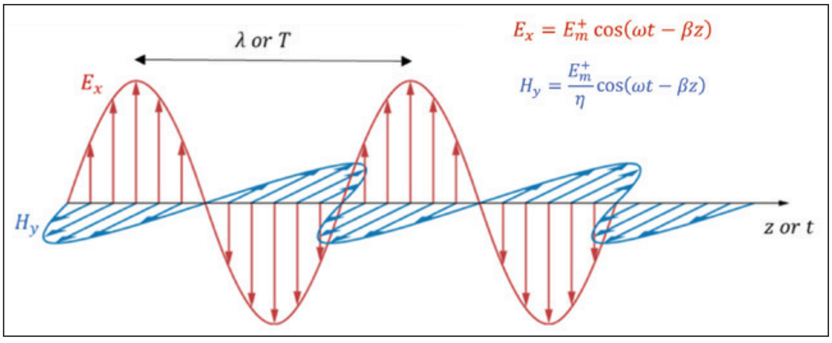

Figure 3: Sinusoidal EM wave in a lossless medium

The wavelength λ is related to the velocity of propagation and frequency by

$$\lambda = \frac{v}{f} \tag{3.12}$$

The phase constant β is related to λ by

$$\beta = \frac{2\pi}{\lambda} \tag{3.13}$$

From Eq. (3.13) we obtain

$$\beta z = 2\pi \frac{z}{\lambda} \tag{3.14}$$

We refer to *z* as a physical length and to *z*/λ as the *electrical length*. The time domain solutions in Equations (3.10) can now be written in terms of the electrical length as

$$E_x = E_m^+ \cos\left(\omega t - 2\pi \frac{z}{\lambda}\right) + E_m^- \cos\left(\omega t + 2\pi \frac{z}{\lambda}\right) \tag{3.15a}$$

$$H_y = \frac{E_m^+}{\eta} cos\left(\omega t - 2\pi \frac{z}{\lambda}\right) - \frac{E_m^-}{\eta} cos\left(\omega t + 2\pi \frac{z}{\lambda}\right) \quad (3.15b)$$

The definition of electrical length leads to the concept of the *electrically short structures*.

Physical object is electrically short if its electrical length z/l ≤ 1/10 *or equivalently if its physical length* z ≤ l/10.

If the physical object is electrically short, then the lumped-parameter circuit models are an adequate representation of that object. This also means that we can use Kirchhoff's laws instead of Maxwell's equations to analyze the circuit models.

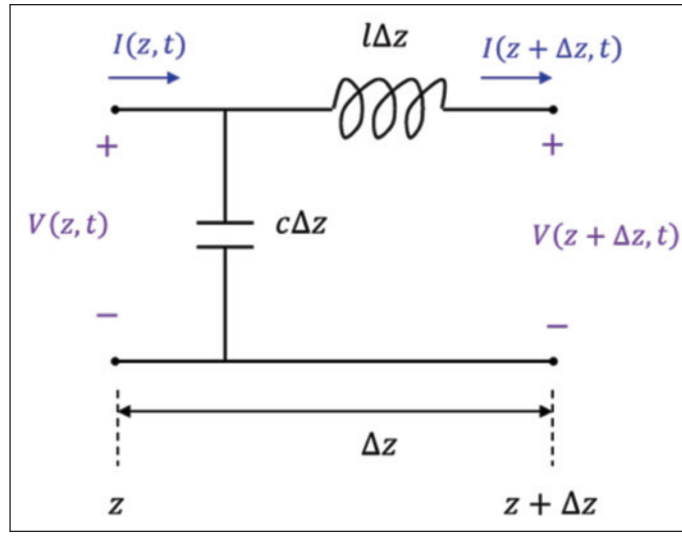

Figure 4: Single segment of a lossless transmission line

4. VOLTAGE AND CURRENT WAVES ALONG A TRANSMISSION LINE

In this section, we show that the voltages and current signals propagate as waves along a transmission line.

To obtain the transmission line equations, let's consider a single segment of a lossless transmission line shown in Figure 4.

The distributed parameters describing the transmission line are: *l* – inductance per-unit-length (*H/m*) and *c* – capacitance per-unit-length (*F/m*).

Writing Kirchhoff's voltage law around the outside loop results in

$$-V(z,t) + l\Delta z \frac{\partial I(z,t)}{\partial t} + V(z+\Delta z,t) = 0 \quad (4.1)$$

or

$$V(z+\Delta z,t) - V(z,t) = -l\Delta z \frac{\partial I(z,t)}{\partial t} \quad (4.2)$$

Dividing both sides by Δ*z* and taking the limit gives

$$\lim_{\Delta z \to 0} \frac{V(z+\Delta z,t) - V(z,t)}{\Delta z} = -l \frac{\partial I(z,t)}{\partial t} \quad (4.3)$$

or

$$\frac{\partial V(z,t)}{\partial z} = -l \frac{\partial I(z,t)}{\partial t} \quad (4.4)$$

Writing Kirchhoff's current law at the upper node of the capacitor results in

$$I(z,t) = I(z+\Delta z,t) + c\Delta z \frac{\partial V(z+\Delta z,t)}{\partial t} \quad (4.5)$$

or

$$I(z+\Delta z,t) - I(z,t) = -c\Delta z \frac{\partial V(z+\Delta z,t)}{\partial t} \quad (4.6)$$

Dividing both sides by Δ*z* and taking the limit gives

$$\lim_{\Delta z \to 0} \frac{I(z+\Delta z,t) - I(z,t)}{\Delta z} = -\lim_{\Delta z \to 0} c \frac{\partial V(z+\Delta z,t)}{\partial t} \quad (4.7)$$

or

$$\frac{\partial I(z,t)}{\partial z} = -c \frac{\partial V(z,t)}{\partial t} \quad (4.8)$$

Equations (4.4) and (4.8) constitute a set of first-order coupled transmission line equations. These equations can be decoupled and expressed as [1]:

$$\frac{\partial^2 V(z,t)}{\partial z^2} = lc \frac{\partial^2 V(z,t)}{\partial t^2} \quad (4.9a)$$

$$\frac{\partial^2 I(z,t)}{\partial z^2} = lc \frac{\partial^2 I(z,t)}{\partial t^2} \quad (4.9b)$$

Compare the Equations (4.9) to the Equations (2.5) describing the EM wave. These two sets of equations have the same mathematical form. This means that the solutions of these two sets will have the same mathematical form! This also means that the voltage and current propagate as waves along the transmission lines!

The general solutions to the transmission-line equations (4.9) are [1]:

$$V(z,t) = V^{+}\left(t - \frac{z}{v}\right) + V^{-}\left(t + \frac{z}{v}\right) \tag{4.10a}$$

$$I(z,t) = \frac{1}{Z_C}V^{+}\left(t - \frac{z}{v}\right) - \frac{1}{Z_C}V^{-}\left(t + \frac{z}{v}\right) \tag{4.10b}$$

Z_C is the characteristic impedance of the line

$$Z_C = \sqrt{\frac{l}{c}} \tag{4.11}$$

The function $V^{+}(t - z/v)$ represents a forward-traveling voltage wave traveling in the +z direction, while the function $V^{-}(t + z/v)$ represents a backward-traveling voltage wave traveling in the -z direction.

Similar statements are valid for the current waves. The total solution consists of the sum of forward-traveling and backward-traveling waves.

The velocity of the wave propagation along the line is given by

$$v = \frac{1}{\sqrt{lc}} \tag{4.12}$$

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PRODUCT Showcase

ADVANCES IN CMOS TECHNOLOGIES LEADING TO LOWER CDM TARGET LEVELS

By Mujahid Muhammad and Robert Gauthier for EOS/ESD Association, Inc.

Can you continue aiming for typical CDM protection levels?

Mujahid Muhammad is an ESD Design Engineer at GLOBALFOUNDRIES with more than 20 years of experience in the field of ESD and Latchup. He presently leads a group of industry wide ESD experts on updating the Industry Council White Paper 2 on CDM Targets and eventually the JEDEC JEP157 standard.

Robert Gauthier has led the worldwide ESD/latchup team within GlobalFoundries. He is currently on the ESDA Board of Directors and an active member on the ESDA EXCOM team. He was one of the founders of the International ESD Workshop (IEW) and is a former General Chair of the EOS/ESD Symposium.

INTRODUCTION

The ESD Design Window (ESD-DW) has been steadily shrinking over time due to technology scaling not only from a smaller feature size but also as the device's architecture has changed over time. Higher speed interfaces are driving the need for lower capacitive loading and higher on-resistance ESD devices. As package sizes increase, peak CDM currents increase as well, putting additional pressure on making improved ESD devices in each new technology generation. As technologies scale, metal resistance at lower metal levels continues to increase, which contributes to increasing the clamping voltages within ESD devices.

The above factors contribute to significant challenges in meeting the generally accepted CDM targets of 250V or 500V. Meeting functional performance and CDM existing targets can be nearly impossible for some types of I/O interfaces depending upon the type of circuit topology used in each design. In this article, we will describe the ESD-DW and the reasons for its continued shrinking. These effects mentioned are summarized in Figure 1 are driving the need for lower CDM targets

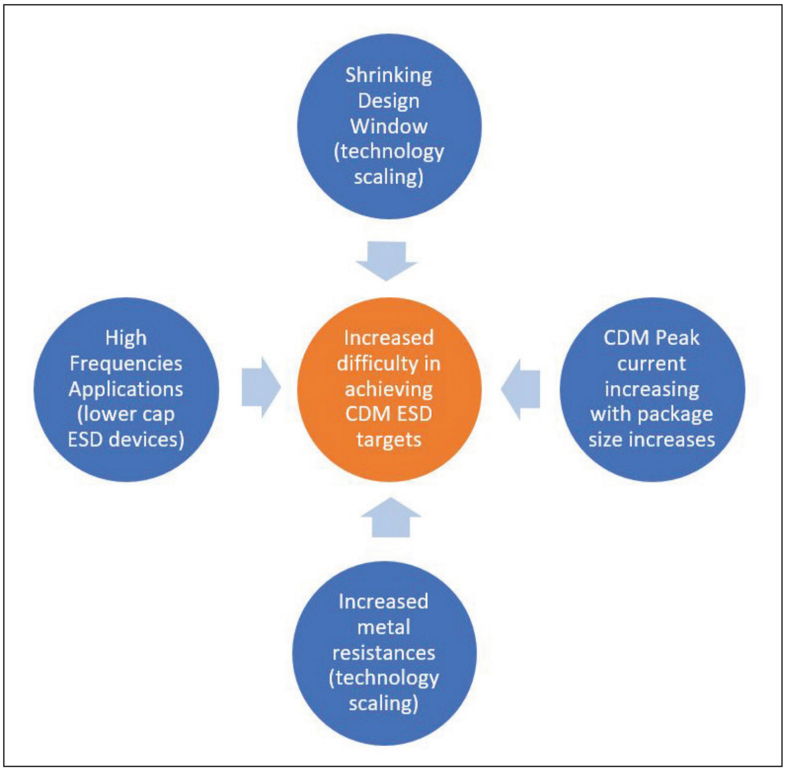

Figure 1: Factors driving the need for lower CDM ESD targets

GOAL OF AN ESD PROTECTION NETWORK

The key goal of the ESD protection device/network on any semiconductor integrated circuit (IC) product is to ensure if an ESD event does occur to any pin of the IC, that pin does not get damaged. It is necessary that any robust ESD protection device added to the pin can handle and divert the ESD event current away from the sensitive internal circuits of the IC. A basic ESD protection network using ESD elements (such as diodes) and Power clamp is shown in Figure 2.

In addition, the overall ESD protection methodology needs to ensure that during an ESD discharge, the protected internal circuit devices should not breakdown due to the increased voltage drops across the parallel ESD devices.

ESD-DW – HOW IS IT DETERMINED?

The ESD-DW is defined as the region between the IC operating voltage (Vop) and the circuit breakdown voltage (Vbd). The ESD-DW shown in Figure 3 defines a visual representation of the voltage and current range, where the ESD protection devices need to operate to protect the I/O and supply interfaces.

ESD protection devices should not turn on during the normal IC functioning voltage range (IC Operating Area), nor should they operate in the IC Reliability Constraints region beyond the internal circuit breakdown voltage (Vbd) during an ESD event. The ESD-DW (Vbd – Vop) has shrunk significantly as technologies scale due to the reduction of the output device trigger voltages and receiver device oxide breakdown voltages [1]. This is quantitatively highlighted in Figure 4, which shows how the ESD-DW has scaled, going from 350nm down to 12nm technology nodes. The design window has shrunk by approximately 65% scaling from 350nm down to 12nm.

WHAT ARE THE NEW TECHNOLOGY ADVANCES DRIVING THE LOWERING OF ESD CDM TARGETS?

Technology scaling has resulted in the IC operating voltage slowly reducing and flattening out in the 0.5 to 1.2 volt range [2]. It has also resulted in a reduction of MOSFET drain/source breakdown (bipolar turn-on) and oxide breakdown voltage under ESD conditions. Specifically, the reduction of the driver device trigger voltages are driven by a decrease in effective gate length, while the reduction in receiver device oxide breakdown voltages are

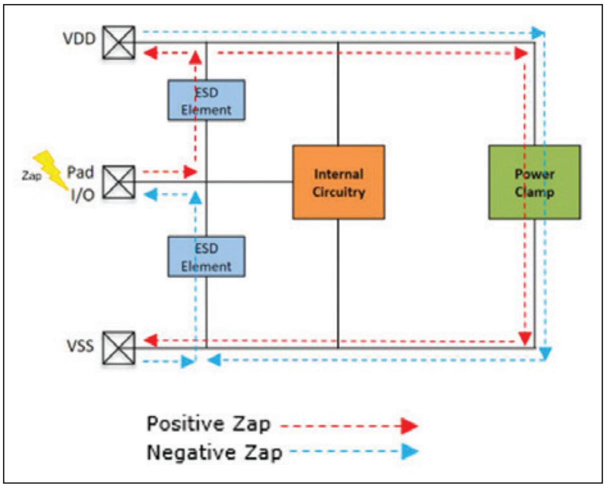

Figure 2: Basic circuit topology of an ESD protection network

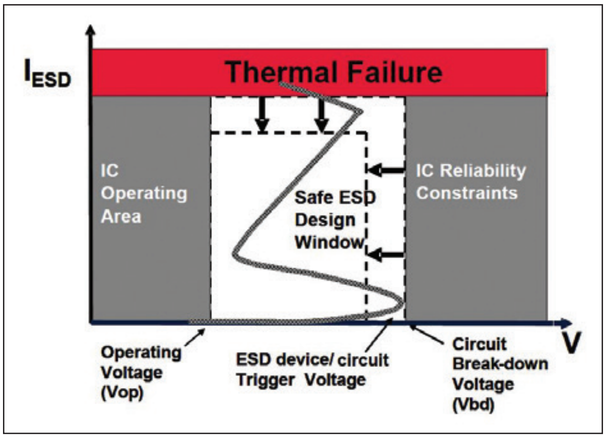

Figure 3: ESD-DW Visual Representation

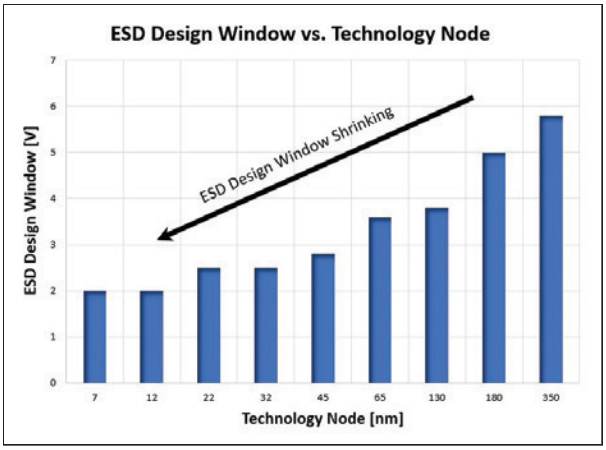

Figure 4: ESD-DW versus Technology Node.

driven by the decreasing gate oxide thickness. These trends are shown for the thin-oxide NFET devices in Figures 5 and 6.

The decreasing metal interconnect thickness due to technology scaling has also resulted in increased lower metal level resistances, such as shown in Figure 7 for first level metal leading to higher potential wiring resistances within ESD devices, especially in the design of low capacitance ESD devices with a limited number of metal levels used to reduce back end of line capacitance in these devices. Figure 7 shows the Metal 1 sheet resistance ratio trend in multiple technologies relative to its value in a 180nm technology. For example, the Metal 1 resistance in 45nm is about 18X the value in 180nm. The temperature increase with this resistance increase, and reduced power handling of the device also leads to the thermal (y-axis) shrinking of the ESD-DW.

WHAT ARE THE TRENDS IN PACKAGE SIZES?

The increasing trend for more computing power has led to increasing package pin count and overall package size. Figure 8 shows this trend. Increasing package sizes lead to higher CDM peak current, as shown in Figure 9. These trends are leading to further challenges to achieving CDM ESD targets [3], [4].

WHAT ARE THE TRENDS IN HIGH-SPEED INTERFACE DATA RATES?

Figure 10 shows the trend of increasing High-Speed Serial (HSS) link data rates with technology scaling. Increasing data rates requires a reduction in the ESD device capacitance, typically resulting in smaller ESD device areas. This is in contrast to the previously mentioned need for larger ESD devices to handle larger CDM current when package sizes increase.

WHAT ARE THE NEXT STEPS IN CDM TARGET REDUCTIONS?

The impact of the ESD-DW reduction applies to any type of I/O ESD protection strategy but is most critical for ICs that have ultra-high-speed interfaces in advanced technologies, which are often packaged in larger packages. These high-speed interfaces

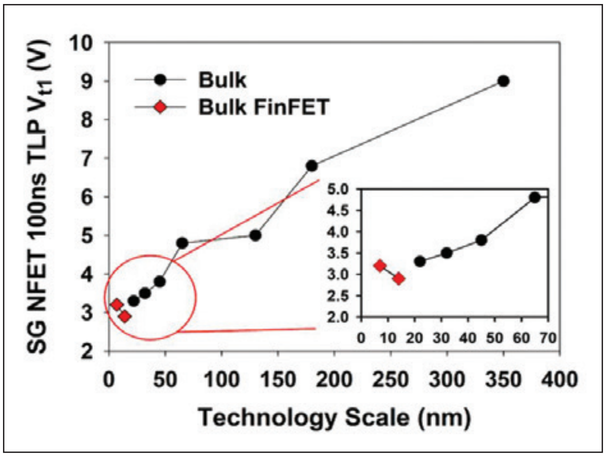

Figure 5: Thin-oxide (SG) NFET trigger voltage Vt1 vs. Technology Node

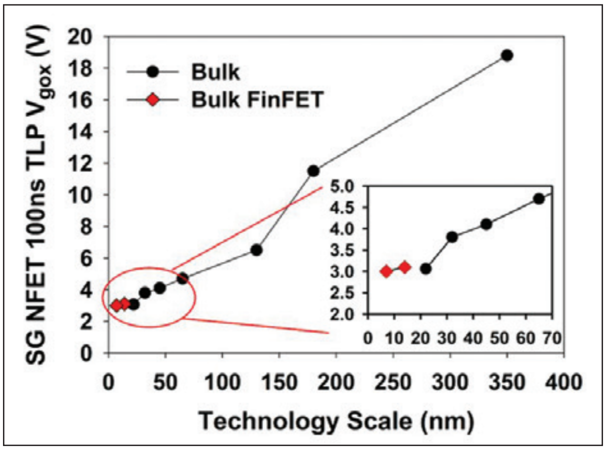

Figure 6: Thin oxide (SG) NFET Oxide Failure Voltage (Vgox) vs. Technology Node

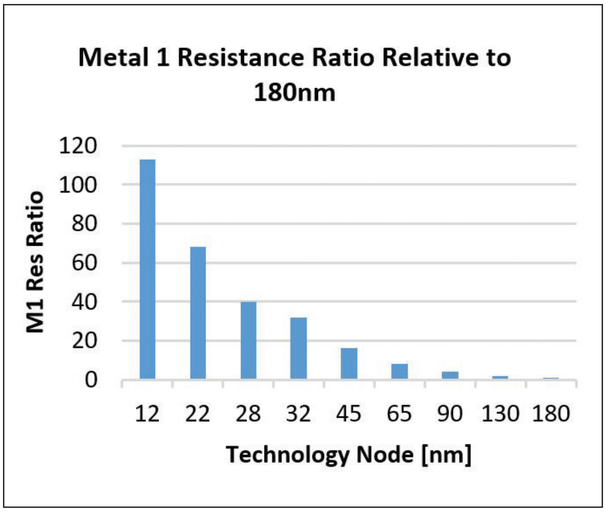

Figure 7: Metal 1 resistance ratio relative to 180nm vs. Technology Node

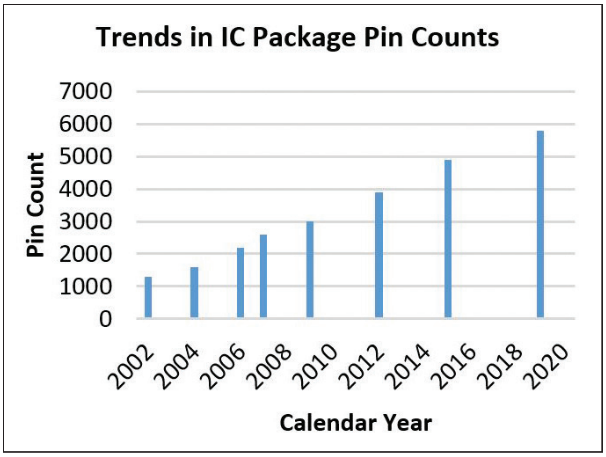

Figure 8: Package pin count vs. year

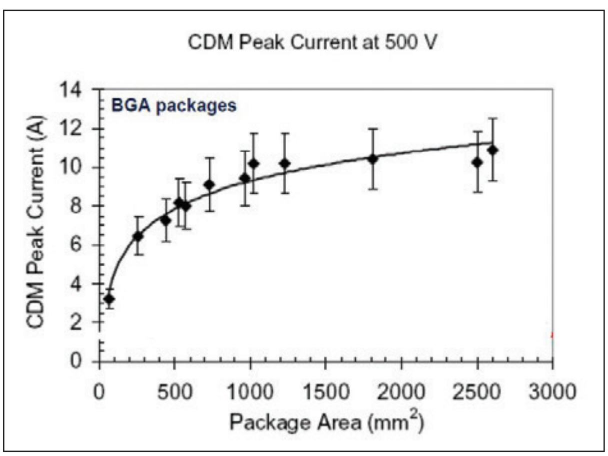

Figure 9: Peak Current vs. Package Size

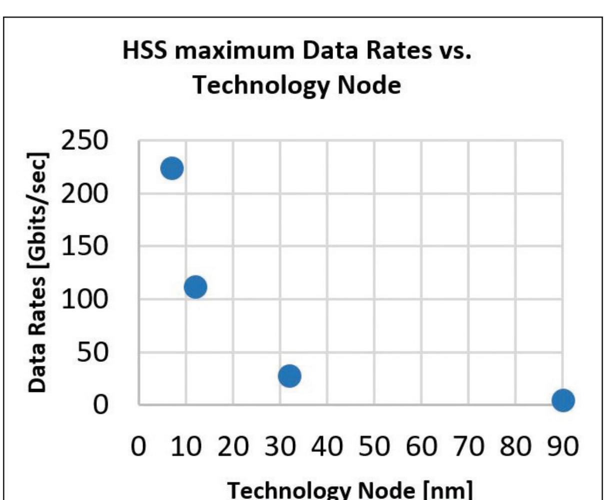

Figure 10: HSS maximum data rates vs. Technology Node

most always use thin-oxide devices, where the need for high-speed performance does not allow for the addition of secondary ESD protection devices. In fact, high-speed interfaces have the additional challenge of requiring reduced ESD device capacitance. Larger package size ICs lead to larger CDM currents for a given CDM voltage (up to 2A to 2.5A per 100 volts of CDM voltage). In addition to larger CDM currents, in advanced technologies like 7nm, the V_{t1} trigger voltage of thin-oxide driver devices is as low as 3.2V. Given the driving factors discussed as technologies scale (smaller ESD-DW, lower capacitance budget for ESD, higher metal Rs at lower metal levels, larger package sizes leading to higher peak currents), a realistic target coming on the horizon is in the 125-150 volt range for CDM target levels for high-speed/RF interfaces in advanced technologies. This reduced CDM target is currently under development and will be presented in more detail in the upcoming revision of the ESD Industry Council White Paper 2 on CDM [4].

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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.

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April 12
Introduction to Measurement Uncertainty

April 13-14
Advanced Printed Circuit Board Design for EMC+SI

April 13-14
Mechanical Design for EMC

April 13-16
Applying Practical EMI Design & Troubleshooting Techniques

April 22-23
Principles of Electromagnetic Compatibility

May 11
Annual Chicago IEEE EMC MiniSymposium

May 11-14
Applying Practical EMI Design & Troubleshooting Techniques

May 13
EMC Fest 2021

May 17-20
IEEE International Instrumentation & Measurement Technology Conference

May 19-20
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Due to COVID-19 concerns, events may be postponed. Please check the event website for current information.