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FCC Looks for "Fresh Approach" in Combatting Robocalls

As part of its ongoing effort to stem the rise in illegal robocalls, the U.S. Federal Communications Commission (FCC) is now exploring options to block robocalls along every point in the call path.

In a Notice of Proposed Rulemaking (NPRM), the Commission is proposing rule changes that would require service providers to give consumers accurate caller name and other information to minimize the challenge of evaluating calls originating from unknown sources. Additional changes would require originating providers to verify that the caller's name and other identification information are accurate and secure, thereby supporting consumer trust and confidence.

The NRPM also details potential steps to reduce the number of scam calls originating from outside the U.S., including advanced call blocking analytics, so that consumers can identify when a call is originating from a foreign country.

The FCC's latest proposal is intended to build on the STIR/SHAKEN caller ID authentication framework, developed by industry to support the goals of the federal Truth in Caller ID Act of 2009.

The Growth of AI Aligns with E-Waste

The rapidly increasing use of technologies based on artificial intelligence (AI) is bringing potentially life-changing benefits to our commercial and industrial industries, as well as to people around the world. However, the promise of AI often clouds our understanding of the impact that AI-driven technologies are having on the environment.

These consequences are highlighted in a recent article posted to the website of *The AI Journal*. Titled "The Energy Crisis Limiting AI's Promise: Hidden E-Waste Explosion Ahead," the article underscores the often-overlooked aspect of AI technology, that is, its dependence on our current energy infrastructure. The article argues that the continued growth of AI will not only overwhelm our energy infrastructure capacity but will also have a disastrous impact on the increase in electronic waste (e-waste).

According to the article, the deployment and use of AI models

currently consume an estimated 415 TWh annually. This is the equivalent of 1.5% of global electricity use today, a usage that is expected to at least double by the year 2030. But, beyond data centers, AI is dependent on data generated by devices and sensors leveraging the Internet of Things (IoT) networks. The number of IoT devices is expected to reach 29 billion (yes, that's "b" as in "billion"!) by the year 2030.

Many IoT devices have a limited shelf life, contributing to the growth in global e-waste. But even more problematic is their current dependence on disposable batteries as an energy source. The article estimates that over 10 billion disposable batteries are produced annually, but that less than 5% of batteries in use are recycled.

This "perfect storm" will likely lead to an estimated 82 million tons of e-waste generated by 2030, not only further impacting the global environment but also constraining the growth of the technology infrastructure needed to support the future growth and deployment of AI technologies.

To offer a light of hope, the article recommends expanding the use of RF wireless power technologies to support increased energy efficiency in AI operations while also addressing growing sustainability challenges. The broader deployment of RF wireless power technology, the article argues, could be the solution to address these and other concerns while eliminating potential barriers to the future growth of AI.

UConn Researchers Investigating Ways to Reduce EM Wave Radiation

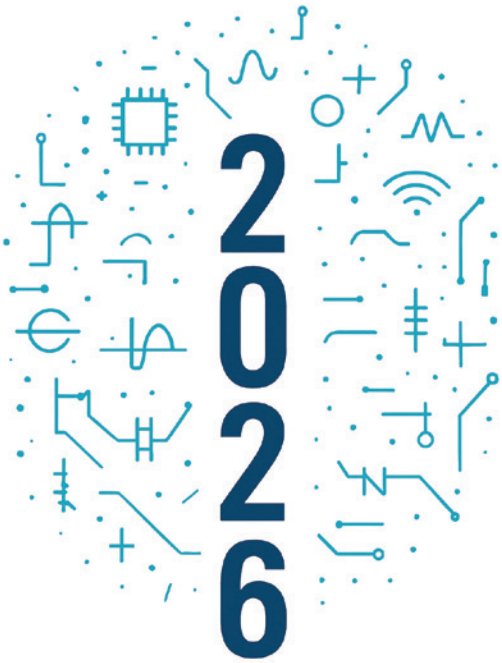
Researchers at the College of Engineering at the University of Connecticut (UConn) are actively working to better understand how different wave-absorbent materials can be used to reduce high-level exposure to electromagnetic (EM) wave radiation.

The research on reducing EM wave radiation exposure is highlighted in an article published in mid-October on the website of "UConn Today." According to the article, Julia Valla, an associate professor in the chemical and biomolecular engineering department, along with her team of students, is evaluating how the use of ion-exchanged zeolites can support the absorption of EM wave radiation.

Under the partial sponsorship of the Air Force Office of Scientific Research, Valla is evaluating the structures of four separate zeolites, including Faujasite, Mordenite, Mordenite Framework Inverted, and Linde Type A, and assessing each structure for its absorption, reflection, and transmission characteristics. Valla and her team are then making small changes to each structure in order to improve its performance.

According to Valla, the objective behind the research is to identify the zeolite structure (or structures) that are most effective in absorbing EM wave radiation. Not only would the findings potentially help to reduce the potential impact of EM wave radiation exposure on humans, but they might also be used to provide additional radiation protection for military aircraft and naval vessels.

Not only would the findings potentially help to reduce the potential impact of EM wave radiation exposure on humans, but they might also be used to provide additional radiation protection for military aircraft and naval vessels.



EXPERT PERSPECTIVES

on the Future of EMC

Where is EMC heading in 2026?

We asked leading experts across the industry to share their vision for the year ahead.

Their perspectives reveal a field in transition—addressing new challenges, emerging opportunities, and the innovations that will shape electromagnetic compatibility.

Bruce Archambeault
MST EMC Lab

Demand for connectivity has grown and will continue to grow as we find more and more uses for the internet. This means that high speed signals will dominate the EMC performance of most EUTs. Signal integrity is no longer separate from EMC, since good EMC performance is critical to good signal integrity, and good signal integrity is critical to EMC performance. Simulation software will continue to be a critical part of the design process. But, simulation/modeling will not take the place of measurement for quite a long time.

Today's computers are powerful, but the technology will need to combine simulation/modeling with artificial intelligence (AI) and machine learning (ML) before simulation software can truly give a pass/fail prediction. Then, possibly, the amount of testing can be reduced (but not eliminated).

Karen Burnham
EMC United

I see two challenges for EMC in the near future: the proliferation of power electronics related to renewable energy, and the expansion of regulatory testing up to 6 GHz. All the inverters and DC/DC converters needed to tie solar and wind power into the grid will continue to be a challenge to control to protect the lower frequency range of electromagnetic spectrum.

On the higher end, testing more products up to 6 GHz (those which may previously have only been tested to 1 GHz) will uncover new problems that need to be addressed.

Getting teams trained to be aware and proactive about EMC concerns will be more important than ever.

Daniel Hoolihan
Hoolihan EMC Consulting

Higher frequencies are the future for EMC testing labs. The spectrum has room at the upper frequencies and the electronic technology is rapidly approaching the highest frequencies for throughput and other reasons.

Currently, the testing lab industry is leading studies of this frequency expansion, a process complicated by the need to develop standard test methods which can encompass the ultra-high frequencies from 30 GHz to 300 GHz. Is it time for the reverberation chamber to take its place as the number one EMC test site?

Regulatory authorities such as the U.S. Federal Communications Commission are also challenged in their efforts to authorize and regulate the next generation of electronic devices.

Bill Koerner
Keysight Technologies

In mid-2025, the FCC adopted a number of rulings regarding the use of bad labs for the certification of products marketed or sold in the United States. Report and Order FCC 25-27, adopted on May 22, 2025, establishes requirements for Telecommunication Certification Bodies (TCBs), test labs, and test lab accrediting bodies, and changed the focus from technical competence to trustworthiness. The focus is on those entities that are located in China, and/or are more than 10% owned by Chinese entities.

We suspect that this will cause a major shift in testing to other countries, perhaps Southeast Asia or the Americas. Device manufacturers and test labs should prepare for challenges in finding available test labs, and for a large increase in testing opportunities.

Chris Semanson, *Senior Contributer*
Renasas Electronics

EMC will have to evolve with the industry and start to account for cheap, dense, AI-adjacent hardware. Everyone wants flagship features in their widget, which makes coexistence, manufacturability, and test coverage brutally hard on a budget. As connectivity and AI drops into toys, wearables, and white-label IoT, teams won't have HDI, exotic laminates, chamber time, or margin-rich processes, and methods common to premium builds will be absent, ignored or unavailable.

So, while there will always be an emphasis on studying how to properly power large data servers and systems, managing small feature-packed consumer products will be just as challenging. Expect renewed focus on architecting, common-mode control, and radiating interconnects, not after-the-fact ferrite band-aids.

Success will mean passing on four-layer FR-4 with 0201 parts with sub-$1 manufacturing.

Min Zhang, *Senior Contributor*
Mach One Design

As we look toward the next generation of power electronics, the adoption of both GaN and SiC devices is accelerating, creating a challenge between electronics design, thermal management, and electromagnetic compatibility. We are seeing these high-power designs force a critical trade-off between thermal performance and EMC control, with significant common-mode noise emerging across a wide frequency spectrum, often beginning at the lower bounds of emission standards.

I anticipate that the push for smaller form factors will drive the adoption of integrated active filters to minimize component size, while electrostatic shielding between the heatsink and components may become essential for containing high-frequency noise by providing a controlled path for circulating currents.

Success will depend on a co-design methodology where thermal and EMC considerations are addressed concurrently from the earliest stages, moving beyond sequential fixes to a truly integrated system approach.

Can Test and Certification Ever Keep Up With Innovation?

As the IoT world evolves, product developers and innovators are creating an ever-increasing number of use cases for existing products or designing new products to achieve the objective of providing a solution to a real-world problem and using internet-connected technology to do it. Around our homes, we have the ability to control many, if not most, of our household appliances and will use existing standards and regulations to demonstrate compliance with whatever legislative requirements exist, but do these standards and regulations continue to be fit for purpose? How can they ever keep up to date with the latest technology and innovation? This paper explores the topic and highlights how Element continues to play its part in addressing this challenge.

Most countries and regions throughout the world have their own product regulations, helping to keep their citizens safe and reduce dangerous products entering their market. These regulations will typically use test standards as a means of providing evidence and consistency across industries to demonstrate adherence to the regulations, either as National standards or the parent International one.

While regulations are mandatory, manufacturers need to ensure that their products will work in the ecosystem that they are designed to operate – lets face it, if you want a new pair of headphones, you're probably going to look for a Bluetooth logo, or if its directly connected to the internet, a Wi-Fi router, as opposed to any other type of broadband wireless access system. These industry standards are written by a cross section of stakeholders and bring the expertise of IoT innovators, test and certification companies, and protocol experts together to create interoperability standards, thus ensuring the connectivity in the ecosystem.

Accepting that there are two types of test standards and certification, underpinning regulations and IoT innovation, comes an appreciation that regulatory standards have to be written with technology neutrality in mind for regulations, to make to open for all. On the other hand, IoT standards have to be written by a cross-functional group of stakeholders to ensure that, as technology and use cases are designed, there are a series of tests and assessments that are developed in parallel to ensure that the products work and are fit-for-purpose.

It's worth reflecting on an old industry saying, "the beauty of standards is that there are so many to choose from," which seems quite relevant to highlight how difficult it is for companies such as Element to keep up with the latest developments. There are so many different 'alliances' formed to solve a particular industry challenge, which one should it support and participate in? The reality is that no one knows until manufacturers start developing products and consumers start buying them – ultimately, just like Betamax and VHS video tapes of old, there will only be one survivor.

Element focuses on key end markets where it wants to operate and where failure is not an option due to the inherent risks. The ever-increasing complexity of IoT products needs companies like Element to participate in a mixture of regulatory and IoT industry standard bodies, as they have particular experience to offer based on their practical experience of testing.

Element rises to these challenges and leads in a number of key areas of product regulations and standards, from chairing the IEC Conformity Assessment schemes, to participation and leadership roles in ANSI and International standards, to participating in expert working groups of IoT innovation and protocol standards.

The benefit for Element and its customers is not only having a detailed understanding of the requirements, but also the background as to how these have been formulated and why. Element customers then benefit from reduced testing regimes where data to support one certification activity can be used to provide evidence for another, helping to reduce cost and, more importantly, time to market.

IDENTIFYING COUPLING IN REAL WORLD SYSTEMS

A Beginning Guide to How Noise Moves in a System and How to Stem It

Chris Semanson is a Senior Contributor to *In Compliance Magazine*, and is a Senior Staff Engineer in the systems and solutions team at Renesas Electronics. He has software and hardware experience across a wide range of applications in embedded systems from automotive to inverter management at John Deere and Ford Motor Company. Semanson holds a Master's degree in Electrical Engineering from the University of Michigan Dearborn, where he studied under Mark Steffka in both computer and electrical engineering. He can be reached at Chris.Semanson.yf@renesas.com.

By Christopher Semanson

Unintended electromagnetic coupling shows up in more ways than we care to admit. Picture this: your latest PCB assembly comes back from fabrication, you power it up, clip on the scope, load the bring up code, and the screen fills with ragged edges and random spikes. The demo you expected to run clean now looks like a jittery mess, and the system that behaved in simulation suddenly appears unpredictable.

You ask around and hear the classic replies: "That is just noise," or "That's just crosstalk." We have all been there. Those symptoms are almost always attributed to unintentional coupling between circuits, which is energy moving along paths we did not intend to create. The usual suspects are layout choices, component selections, cable and harness effects, or enclosure details that felt minor at the time but that have big consequences during bring up. These decisions generally range from:

- A slightly long return path, a decoupling choice that looks fine on paper; to
- A FET gate loop with more area than it should have can be enough to tip a marginal design negative.

This article builds a practical framework you can use on the bench and in the chamber. We will start with how energy moves through a system: conductively through return paths and impedances, and electromagnetically through capacitive, inductive, and radiated paths. We will then separate problems by mode, common mode versus differential mode; that single classification often points directly to the shortest fix. Along the way. we will introduce and align around a simple vocabulary, such that debugging can center around the EMI model: source, path, and receptor, all without talking past one another.

From there, we will examine the most common conducted mechanisms that show up during bring up, including concrete examples and common remedies, plus share notes on how to verify that the change addressed the issue.

Next, we will step into a near field versus far field coupling. We will discuss what changes in the reactive region, why fixes that work on some issues may not work on others, and how to recognize when a local magnetic or electric field interaction is dominating the behavior.

Finally, we'll lay out a field-tested debug workflow for when someone asks, "Why does the circuit quit working when the motor starts?" The goal is not only to quiet the waveform today, but to give you a

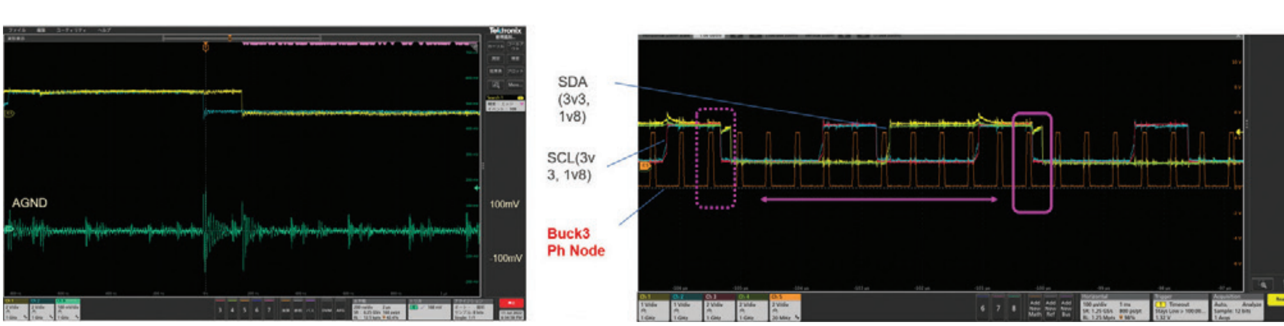


Figure 1: An oscilloscope shot of "noise"

repeatable method to identify the path, identify the mode, and remove the coupling with minimal impact on schedule, so you can get back to your product.

DEFINING COUPLING, AND THE BASIC METHODS FOR ENERGY TRANSFER

When two parts of a system couple, it simply means energy leaves one circuit and shows up in another. The result can be an unwanted perturbation on a signal, elevated emissions, or failure of the intended function. To understand how that energy moves across a PCB, a module, or a full system, we start with two top-level categories:

- *Conducted coupling*: Energy rides on conductors and their return paths. It flows through intentional interconnects and unintended impedances in planes, cables, and grounds.
- *Radiated coupling*: Energy leaves a structure as an electromagnetic field and is received by another structure, much like an antenna.

And your problems rarely live in a tidy box, and this results in a spectrum of coupling. Conducted issues travel along wires and planes and through the parasitics that you did not mean to build. Radiated issues depend on the radiating structure, the wavelength, and which field is dominant at the distance of interest. In practice, local capacitive and inductive interactions live in the near field, while true free space wave radiation dominates longer distances. The center of this picture is the crossover (see Figure 2).

A conducted problem can be the source of a common mode noise onto a cable that radiates strongly. A strong external field can induce a voltage or current on a trace that looks exactly like conducted interference at the receiver.

The first step in creating our mental scaffolding is to use the EMC/EMI model, and determine following:

- *The Source* of the coupling, often a signal with a fast edge rate such as a PWM drive signal or communications bus.
- *The Path*, or how the energy is coupling from the source to the receiver.
- *The Receiver*, which is often easily identified as it's the device that is not delivering its intended function.

Taking this mental model into consideration, we'll start by examining the path and focus first on conducted coupling by classifying coupled noise currents as common mode or differential mode. That single decision often narrows the root cause and points to the lowest effort fix.

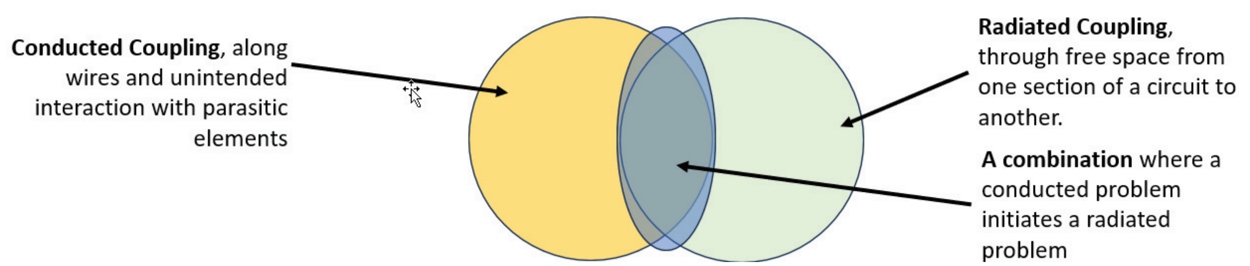


Figure 2: The vein diagram of coupling

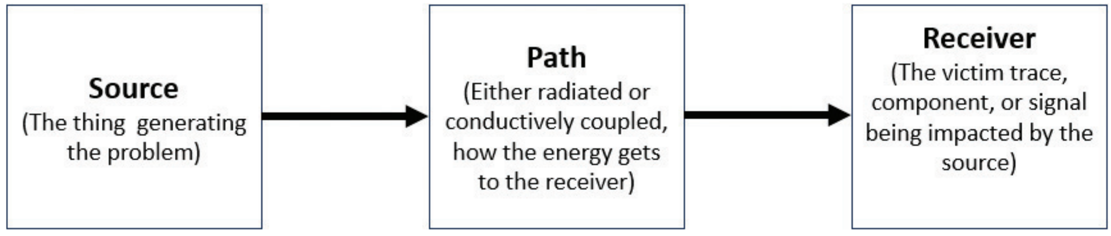


Figure 3: The trust source-path-receiver model

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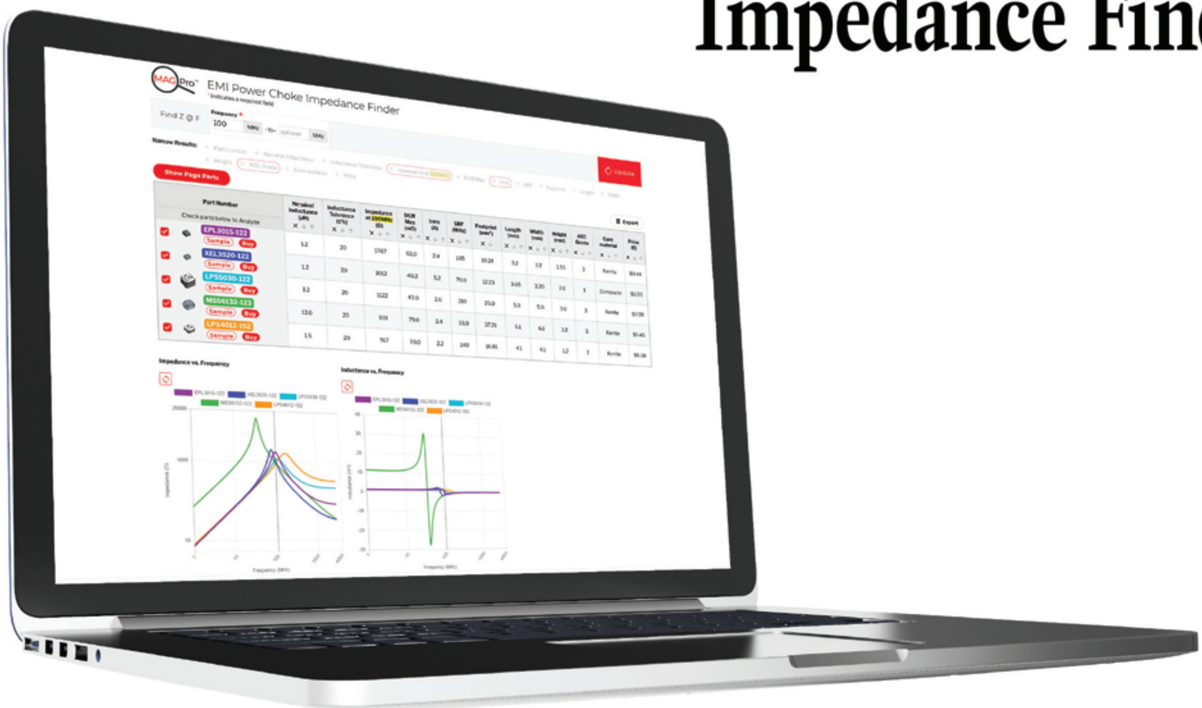

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Conducted Coupling: Differential Mode

Conducted coupling in a system shows up through two main mechanisms, differential mode and common mode. In differential mode, the intended current flows out and back on a loop between a source and a load. Any unintended impedance in that loop turns that current into an unwanted voltage. In common mode, the two conductors share a noise current in the same direction with respect to a reference.

We will start with differential mode because it is the easiest to see and the fastest to fix during startup.

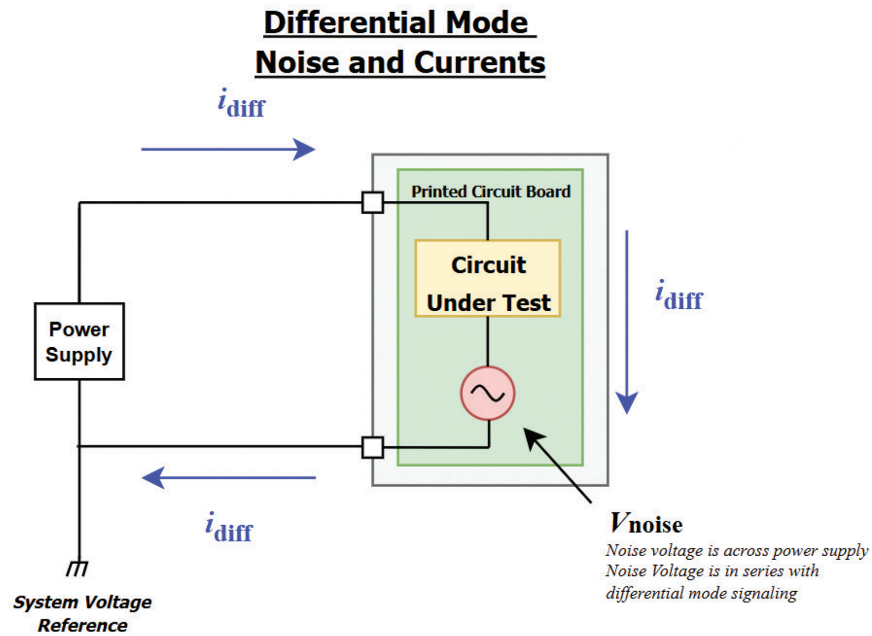


Figure 4: The differential mode current model

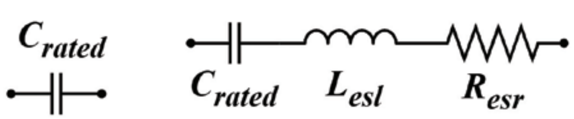


Figure 5: The parasitics of a capacitor

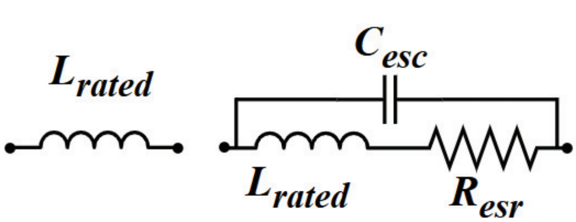


Figure 6: The parasitics of an inductor

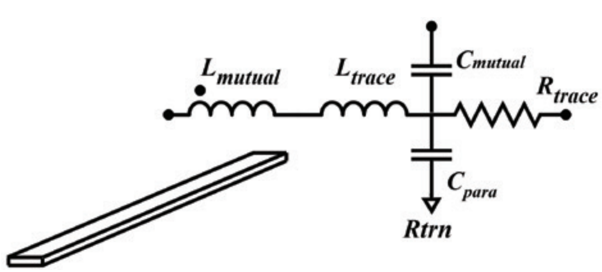


Figure 7: The parasitics of a trace

In Figure 4, the intended operating current is labeled i_{diff}. It leaves the power supply, travels through the circuit under test, and returns to the source on its defined return path. Along the way, the loop encounters small but very real impedances in components, vias, traces, planes, and connectors.

When i_{diff} passes through those impedances, it develops a noise voltage:

- For *resistive* elements, the relationship is an IR drop, $v_{noise} \approx i_{diffcurrent}R$
- For *inductive* elements, the relationship is $v_{noise} \approx L(\frac{di_{diff}}{dt})$
- For *capacitive* elements, the current that must be the source is $i_{sourced} \approx C\left(\frac{dv_{driving}}{dt}\right)$, which shows up as extra ripple or ringing in the loop.

Apart from a basic voltage drop, the main takeaways are:

- The faster you drive current through a series impedance, the greater a voltage error, and
- The larger the loop geometry, the more susceptibility is created in nearby fields.

Hidden Parasitics

Applying these concepts, we can come up with a list of common sources with parasitics you did not plan for in the intended current path:

- *Capacitor parasitics*: The wrong dielectric or package, long pads, or tall stacks push up ESR and ESL. A single X7R 0402 may have about half a nanohenry of ESL; at 100 MHz, that is a few tenths of an ohm, resulting in noticeable ripple. Poor placement lengthens the loop from pin to capacitor and back, which adds trace and via inductance on top of the part's own ESL.
- *Inductor parasitics:* DC resistance (DCR) of an inductor creates a drop under load. Leakage inductance and interwinding capacitance form resonances with surrounding caps. The wrong core material or a part driven into saturation increases effective impedance and injects sharp edges into the loop.
- *Trace and loop parasitics*: Narrow neckdowns, long detours, and unnecessary layer changes add resistance and inductance. A via is on the order of a nanohenry. A few millimeters of thin trace can add several nanohenries. With a 1-amp step in 50 nanoseconds, 10 nH makes about 0.2 V of unwanted drop. Split planes or slots that force the return to wander also enlarge the loop and raise the series impedance that the current must cross. Additionally, close routing and high impedances/low currents can result in mutual capacitive coupling both to return and to other traces, injecting noise currents on victim lines.
- *Connectors and cables*: Pin assignment, contact resistance, and pin inductance add series impedance. A single, fast, high-current pin next to a sensitive analog pin is a prime example of this.
- *Measurement setup*: Long ground leads on a scope probe add inductance and will show ringing that is not on the circuit. The probe can easily become the dominant parasitic in a sensitive loop.

Differential mode coupling is usually correlated with the load current and addressed with decoupling and filtering. You will see supply rails dip or ring when a power stage switches, or a comparator false trigger when a motor phase current steps. The signature often follows the waveform that drives the loop and improves when you shorten the loop or reduce.

Identification and Mitigation of Differential Mode Emissions

The next step in our mental scaffolding is to locate where differential mode emissions arise in a real system. To center the discussion, consider a synchronous buck converter as shown in Figure 8. The goal is to translate an acceptable schematic into a layout that does not amplify parasitics. We will start with a non-optimized layout in Figure 9 and work toward an improved version.

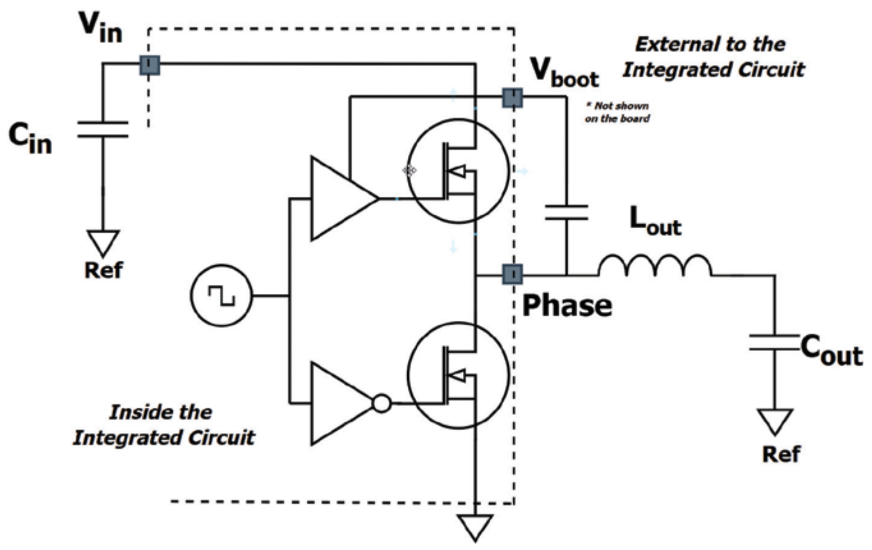


Figure 8: Schematic of a basic synchronous buck converter

When translating from schematic to layout, it's often difficult to balance the design requirements while mitigating differential mode noise.

Begin at the input. The input capacitors provide local DC stability on VIN. Without them supplying the transient current, every switching event pulls current through the upstream harness, and you will see dips on VIN. In the initial layout, the two input capacitors sit at a noticeable distance from the device.

- The fix is to reduce parasitics and shrink the current loop. Use several small case multilayer ceramic capacitors right at the VIN pins and the closest return so the mounting inductance is minimal. Parallel values to spread resonances rather than relying on one part to do all the work.
- Then, shorten and thicken the high current path. Feed VIN with a local plane instead of a long trace, and when a layer change is unavoidable, use multiple vias in parallel to keep the effective inductance low.

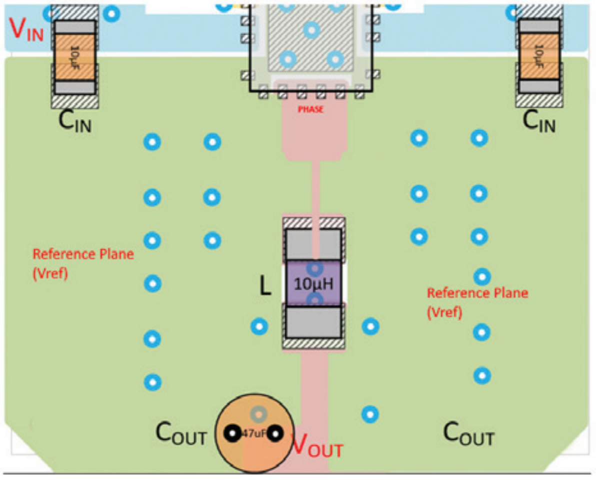


Figure 9: A not-so-great layout of a basic synchronous buck converter

We can realize these two concepts and come up with the following, denoted in Figure 10.

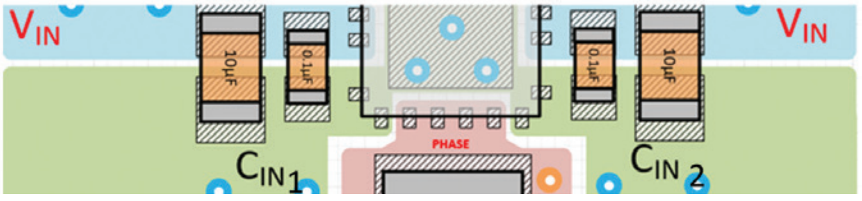


Figure 10: An example of improved input capacitor layout

Now, focus on the switch loop, which is the phase node between the device switching output, the inductor, and the output capacitor bank.

We continue to focus on the copper first, and then move to proper component choices:

- We strive to keep this loop small and planar, making the copper continuous and wide so there are no

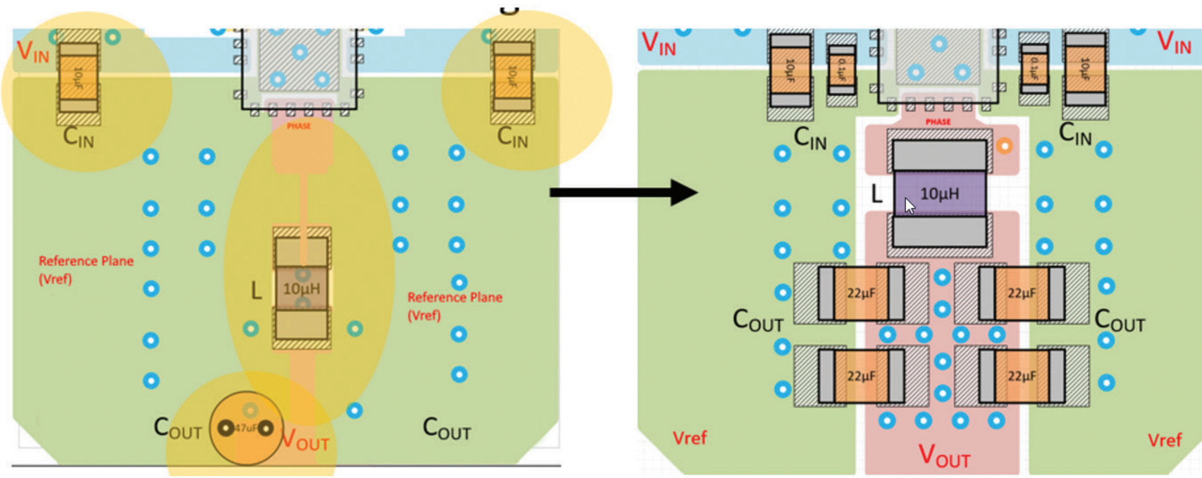


Figure 11: An example of an improved buck layout

skinny neckdowns at device pins or layer transitions that become the dominant impedance. Additionally, we minimize the exposed area of the phase node to reduce its capacitive coupling to nearby nets.

- Place the first output capacitors as close as practical to the inductor and the return so the AC component of the inductor current closes locally.
- Additionally, our focus remains on intelligent component choices to minimize parasitic, an inductor that does not saturate at peak current and whose DCR does not waste margin. Use output capacitors that hold value over bias and temperature. Favor packages and footprints that keep pads short and the current path direct. Shielded inductors and careful body orientation reduce stray fields into sensitive nodes.

As you apply these steps to the non-optimized layout in Figure 10, you will see a pattern emerge. Current loops get shorter and thicker, parasitic inductance drops, and the converter becomes easier to probe and control.

With the differential path under control, the remaining noise that shows up at the chamber often comes from hidden common mode currents traveling along cabling and long traces.

Conducted Coupling: Common Mode

While differential mode emissions usually trace to currents you can point to on the schematic, common mode coupling arises when a signal rides on multiple conductors in the same direction with respect to a shared reference, such as a supply, ground rail over an earth connection. Typical symptoms include:

- Power rail collapse, where a nominally stiff rail shows dips or ripple; and
- Ground bounce, where the reference no longer holds a steady zero volts.

We will use the common mode model in Figure 12.

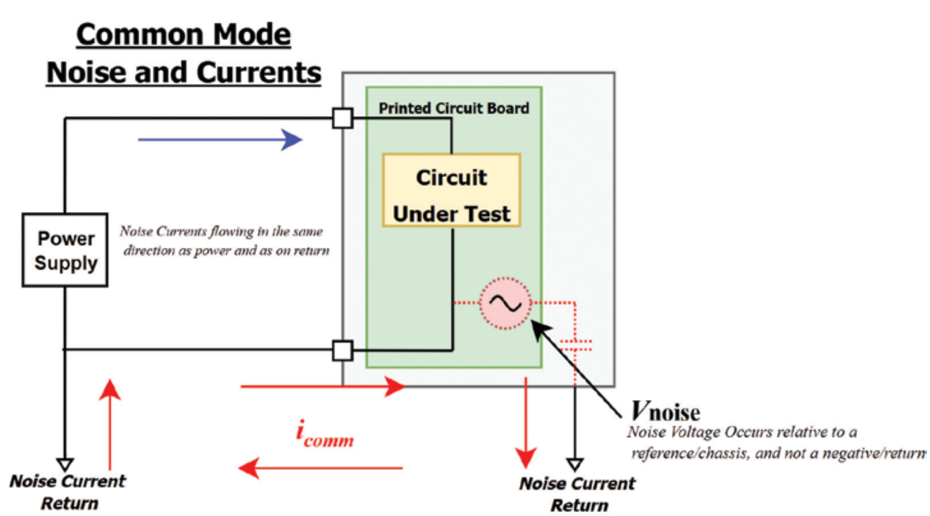


Figure 12: The common mode noise current model

In this model, instead of the functional current shown in blue, a common mode current circulates through the system reference. It often originates from parasitic capacitances between fast switching nodes and nearby structures such as planes, chassis, cables, or the environment. As this current returns to its source, any impedance it meets creates voltage that lifts or disturbs the entire reference.

These currents are sometimes called drift currents or antenna currents. They are tricky because the dominant path is geometric rather than schematic, so the loop may involve the PCB layout, a cable shield, and even the test setup, none of which are obvious in the schematic. As with differential mode, we will discuss the impact of common mode coupling on low voltage signaling in CMOS logic.

Common Mode Coupling Impact on CMOS Logic

Input/output voltage signaling has steadily crept down over time, and I/O levels that were once at 5v are now switching at 1.8v, shrinking noise margins.

That shrinking margin tightens when the receiver decides on a one or a zero. Specifically for CMOS, the input thresholds voltage input high (VIH) and voltage input low (VIL) are defined relative to the local supply and the local reference; this relationship is shown in Figure 13.

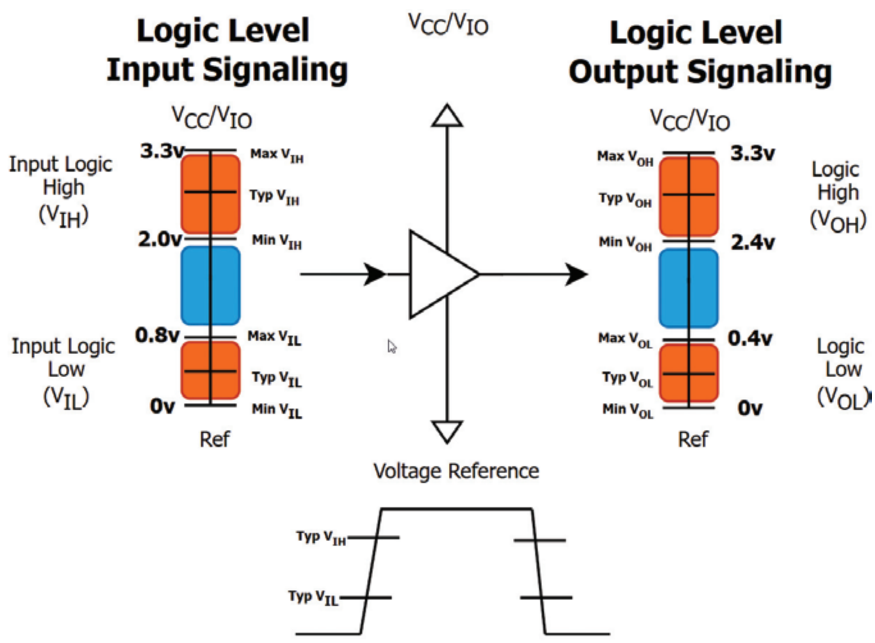


Figure 13: The input output model with basic logic levels for 3v3 logic

This logic only behaves if both the supply rail and the reference stay where we think they are. The reference is easy to overlook on a datasheet because it appears under many names, such as AGND, PGND, GND, or AREF. Inconsistent naming resulting in shared return paths can perturb the reference plane. When the reference lifts, the effective thresholds at the input shift, even though the schematic has not changed.

Figure 14 demonstrates this. The reference moves from zero to a transient positive value. That rise reduces the range where the gate recognizes a valid low. With enough bounce, the receiver can miss a low going transition entirely, which shows up as a missed edge in an SPI or I2C transaction. The same mechanism can also create a false height in a gate driver, as shown in Figure 15.

This is classic common mode behavior. A fast node elsewhere couples through parasitics into planes, cables, or the chassis, launching current that returns through whatever impedance is available. As that current returns, it creates voltage on the shared

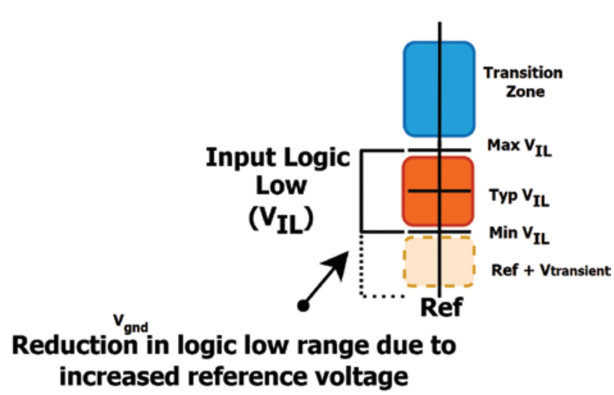


Figure 14: The input low logic level when the reference is lifted by a transient

reference. The receiver does not see the absolute line voltage. It sees line minus reference. If the reference is moving, the apparent level at the pin moves with it. On I2C, where pull ups establish the high level and devices pull down to make a zero, a lifted reference can keep the line above VIL long enough to miss a start, stop, or data bit. On SPI, reference movement can shift both data and clock relative to the threshold, which can violate setup and hold and produce sporadic bit errors; an example of a missed edge is in Figure 16.

Where Does It Come From, and How to Get Rid of It

While differential mode coupling usually comes from non-ideal elements in the intended current path and is often easy to spot, common mode coupling is trickier. You rarely see the dominant path on the schematic. It is set by geometry, by parasitic capacitances from fast signals into planes, chassis, and cables, and by small imbalances that convert differential energy into a net current on a bundle with respect to reference.

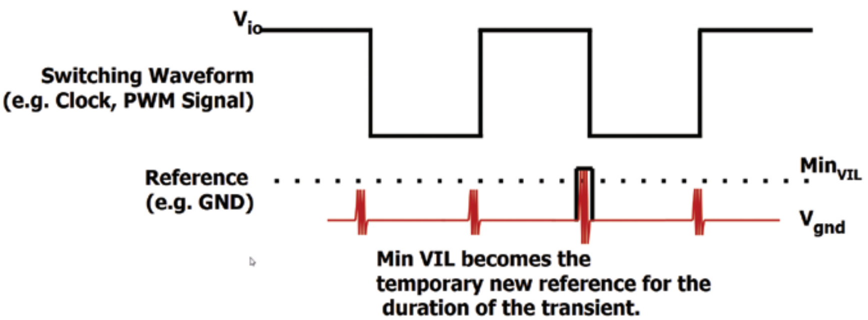


Figure 15: A reference plane being lifted to some minimum level during edge transitions

The most reliable way to keep common mode under control is to bake it into layout and system rules from day one.

This includes keeping return paths contiguous so currents can come back directly under their forward paths. Use zoning and floor planning so high current power stages and sensitive interfaces do not force each other across splits or long detours. Place inputs and outputs so they enter and leave the board without crossing cuts in the reference plane, and avoid plane slots that make the return wander. And in systems with long interconnects, treat the cable interface as its own circuit. Terminate shields with a continuous 360-degree bond at the connector or enclosure, not a long pigtail. Provide a local path that brings common mode current back to its source at the point of entry.

That usually means a combination of a common mode choke in series with the lines and one or more safety-rated Y capacitors from the lines or reference to chassis or a quiet reference. The choke raises impedance to common mode without disturbing the intended differential signal, while the capacitors give the noise a short return.

Balanced routing matters. Differential pairs and matched lines cancel fields and resist conversion to common mode when their impedances and environments are symmetrical. When the pair becomes unbalanced by skew, uneven reference, neckdowns, or a via only on one leg, that cancellation breaks. The remainder becomes common mode that couples to nearby structures and finds a large path through planes or cables.

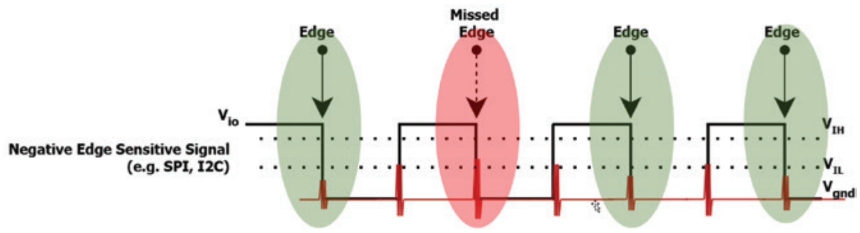


Figure 16: A missed edge on a 12C or SPI trace due to a lifted return plane

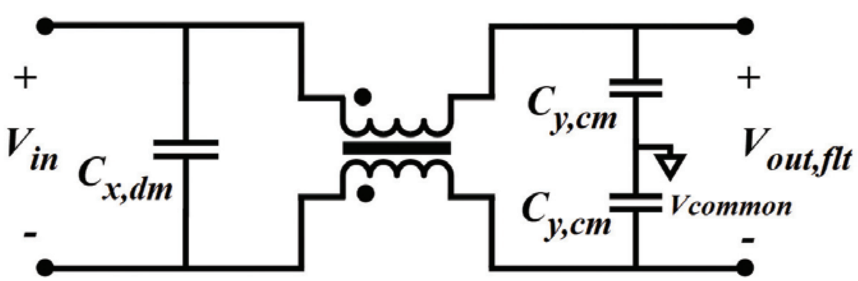


Figure 17: A full line filter with common mode, and differential mode counter measures

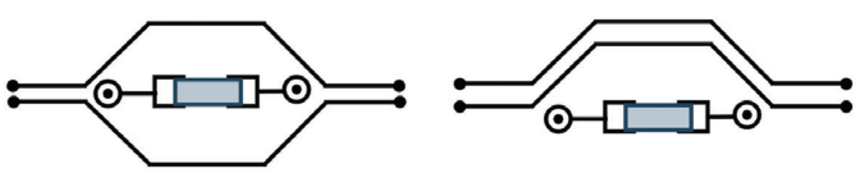


Figure 18: Differential trace routing (left), and an incorrect way that creates imbalanced currents (right), a correction

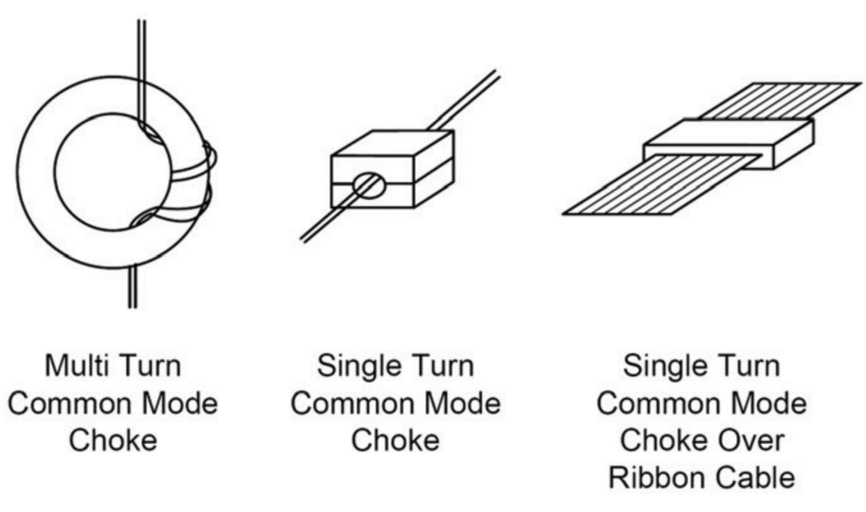


Figure 19: Examples of common mode counter-measures

Filtering for common mode works because it corrects those imbalances and supplies low impedance returns. At the interface, the common pattern is a line filter that includes a common mode choke close to the connector and short capacitive shunts to the reference or chassis on the board side of the choke. Inside the box, stitching capacitors across unavoidable seams or between local and chassis reference at high frequency give noise current a short loop. Where practical, reduce the $\frac{dv}{dt}$ and $\frac{di}{dt}$ of the aggressor with snubbers or control the gate driver parameters so that less energy is driven through parasitic capacitances in the first place.

Identifying the Radiated Nature of a Conducted Coupling Problem

As we move from conducted coupling into radiated effects, remember that many real problems are not clear-cut. Differential and common mode currents that start on copper can excite structures that radiate. Every noise current path can act like an antenna. Our job is to spot it and make that antenna inefficient.

In differential mode radiation, the loop is formed by the intended signal and its return, as shown in Figure 20. The radiated field grows with loop area and with the current edge rate. To manage this:

- At the module level, the remedy is simple in concept: route high current or fast signal loops directly from source to load while keeping a solid, adjacent return so the loop stays small and planar.
- At the system level, use twisted pair or a shielded cable so the forward and return currents stay close together and cancel fields.

Common mode noise is primarily impacted by the length at which currents must travel to return to their source. Here, long conductors, traces, rails, and especially cables behave like monopole or dipole antennas driven by common mode voltage.

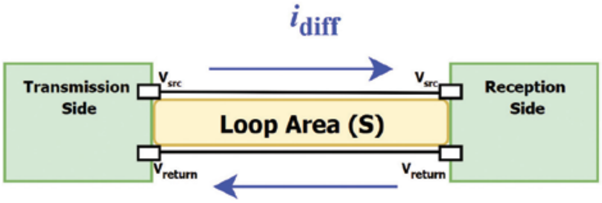


Figure 20: The differential mode model with loop area being the driving factor in its ability to radiate

The strength of the radiation depends on the length of the structure relative to the wavelength and on the impedance of the return through stray capacitances to the chassis and to space, modeled in Figure 21.

Long, unterminated pigtails, breaks in the reference plane, and poor shield terminations make these antennas efficient, and are the low-hanging fruit to be addressed first.

The next section of this article builds on this by mapping the dominant radiated coupling mechanisms and showing how to degrade the antenna that your layout or harness accidentally created.

RADIATED COUPLING MECHANISMS

Radiated coupling is conceptually changing because it asks you to reason about the electric and magnetic fields around a conductor and how those fields change with distance. To help our understanding, imagine you are a neighboring circuit sitting a distance **D** from a source. What you "see" depends on two things: how close you are and the wavelength of the signal.

Up close in the near field, the electric and magnetic parts are not locked together. One often dominates the other, with both appearing out of phase. Energy is mostly stored and returned to the source rather than carried away. Farther out, the fields settle into a traveling wave where the electric and magnetic parts are in phase and tied together, and the ratio **E/H** approaches the impedance of free space (about 377 ohms). Shorter wavelength signals reach this traveling wave behavior at shorter distances. This transition point from the reactive, near field region to far field is thus governed by the following two relationships.

$$D_{\text{distance,meter}} = \frac{1}{10}\lambda_{rise,\,fall} \qquad \lambda = \frac{c}{f}$$

- For small sources where the physical size $D_{src} << \lambda$, the *reactive near field* is roughly within $r < \frac{\lambda}{2\pi}$ (about 0.10-.16 of a wavelength). Beyond that, the fields begin to behave like a radiating wave.

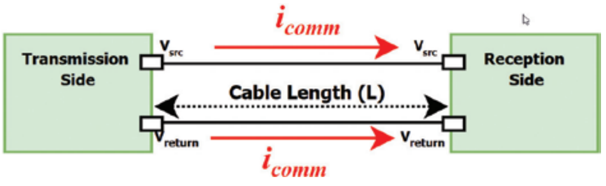


Figure 21: The common mode model shown with cable length being the driving factor in its radiating ability

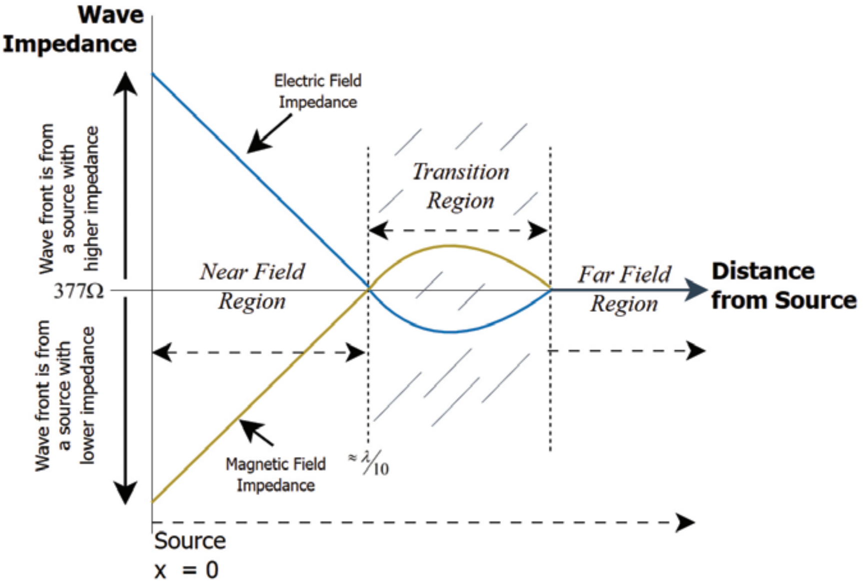


Figure 22: Wavefront as a function of distance from the source, identifying the three impedance "regions" as it traverses to a 377ohm wave.

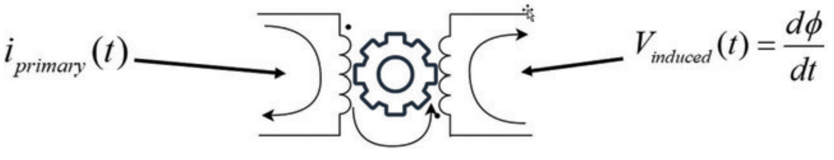


Figure 23: Thinking about inductive coupling as linking two sections of a circuit together via a gear or water wheel

- For larger structures (enclosures, long cables, big boards) the start of the *far field* is better estimated by $r \gtrsim \frac{2Dsrc^2}{\lambda}$. Here, the size of the radiator matters as much as the frequency.

Practically, that gives us a simple relationship shown in Figure 22.

This relationship is described by how close you are to the source. When you are very close to the source, you are in the reactive near field, generated by either:

- *Magnetic sources* (current loops) or
- *Electric sources* (voltage on small structures).

And these tend to couple through your system in different ways. As you step back to distances on the order of a fraction of a wavelength and you enter a radiating field near region where pattern forms but is not yet settled, you're entering what is often called the transition region. Step back farther, and you are in the far field where a simple traveling wave description works, and antenna length and orientation are the dominant coupling mechanisms.

Now we'll focus specifically on near field coupling, defined by either the magnetic or electric field sources.

Near Field Coupling: Magnetic Field

A straightforward way to understand magnetic sources is to use the transformer analogy. Think of the aggressor loop as the primary and the nearby victim loop as the secondary. A transformer works because the two windings are magnetically coupled. Current in the primary creates magnetic flux, and the portion of that flux that links the secondary produces voltage there by induction. Figure 23 shows this relationship.

Think of the inductance that these two entities share as a gear ratio. The size of this coupling gear L_{mutual} grows when the loops have a larger area (bigger gear diameter), sit closer together (tighter mesh), share more overlap and better alignment (more teeth engaged), and when each loop has a tight return path that keeps its area small and well defined.

$$i_{primary}L_{mutual} = \phi \longrightarrow L_{mutual} = \phi / i_{primary}$$

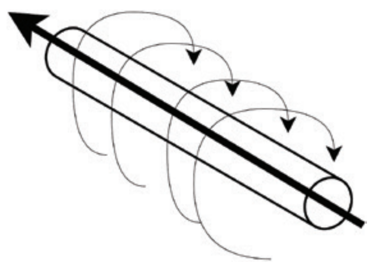


Figure 24: Flux vectors encircling a current flowing down a conductor

Physical Representation

$$\phi = \oint_A B \cdot da$$

$$\phi = BA_{loop}$$

$$V(t) = \frac{d\phi}{dt} = \frac{dB}{dt} A\cos(\theta)$$

$$V(w) = jwBA\cos\theta$$

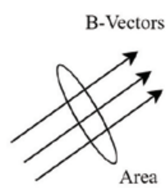


Figure 25: The math that allows you to move from flux vectors to voltage

To understand how stray magnetic fields can couple into, and then induce voltages in, neighboring circuits, we need to better understand the term flux and how it relates to inductance. At a basic level, the amount of linked flux and the current in a circuit are proportional to each other, with inductance as the proportionality constant; this flux is the "stuff" around a current-carrying conductor shown in Figure 24.

We can choose to integrate these flux vectors into an area, as shown in Figure 25.

We then integrate these flux vectors over an area, as shown in Figure 24. If we assume the area stays constant and the time-varying magnetic field creating the flux is roughly uniform across that shape, we arrive at a more intuitive view of *mutual inductance*: how voltage, magnetic field, and area are linked together. This can be summarized with the following analogy:

Envision a net, formed by a neighboring circuit, that catches time-varying magnetic field lines, which, through induction, creates a time-varying voltage in that loop.

You can connect this back to the layout and measurement anytime a loop exists. In Figure 26 on page 26, an inner and outer loop exists, and they're coupled through the shared flux from the source current's magnetic field.

The mutual inductance is shown as linking the source and receiver together, and finally, that induced voltage is then modeled as a frequency dependent voltage source on the receiver circuit or loop.

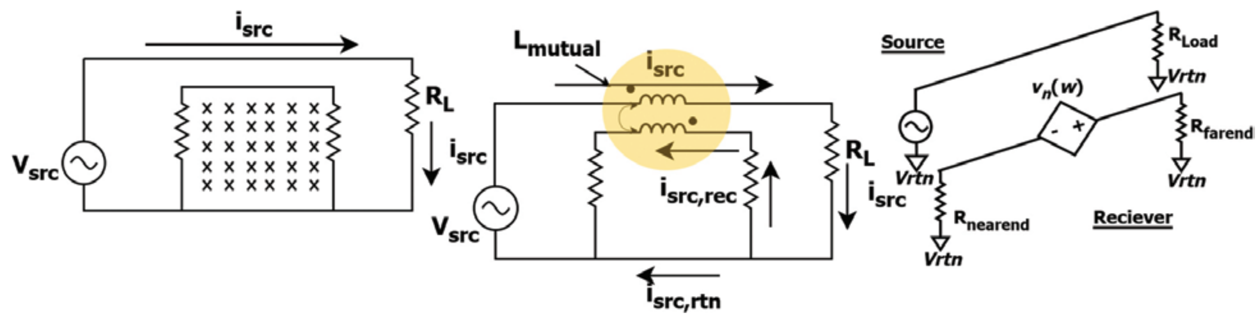


Figure 26: From left to right, the transition from loops in a circuit to the equivalent circuit model showing the source and receiver coupled

Near Field Coupling: Electric Field

Similar to the transformer analogy, near field electric coupling can be modeled as a bad antenna or, more precisely, a small open ended capacitor: one plate with charge separated from another plate by a dielectric. In this picture, the source conductor might be a high-speed digital line and the receiver a sensitive analog trace, with air or solder mask separating them, as shown in Figure 27.

As the electric field from the source extends into space, it forms capacitance to nearby conductors. One end of that capacitance is the source; the other end is the return plane and the neighboring trace or wire that acts as the receiver. This results in coupling between conductors and between each conductor and its return. These parasitic capacitances couple a source to the receiver in differential mode, and they drive the source and its return together in common mode.

The net effect is a current injected into the neighboring circuit, $i_{inj} = \frac{C_m dv_{src}}{dt_{rise(fall)}}$ as shown in Figure 28. The amount of injected current is set by the mutual capacitance and the source slew or frequency content. The resulting voltage at the receiver depends on its input impedance, including any near end or far end termination, and follows the proportionality that $v_{reciever} \approx i_{inj} Z_{in}$.

To control this kind of electric field coupling, we go back to Figure 27 and adjust the few levers we have. Reduce the mutual capacitance by shortening the length that the two conductors run side by side or by increasing their spacing.

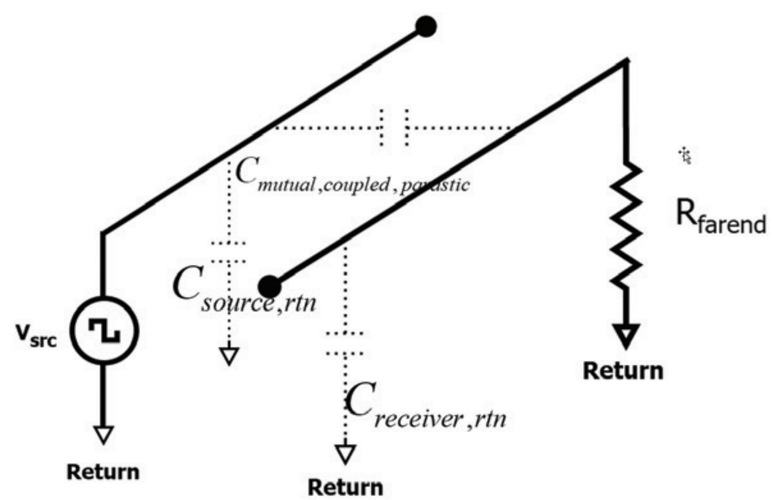
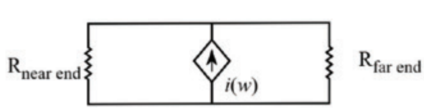


Figure 27: The capacitive coupling model



$$i(\omega) = j\omega R_{load} C_{coupled} V_{source}$$

Figure 28: The equivalent circuit model resulting in an injected frequency-dependent current source in the receiver

$$Cparastic \approx \varepsilon_o \varepsilon_{fr4} \left.\frac{\text{length}}{\text{distance}}\right|_{\substack{l=\text{length of parallel exposure} \\ d=\text{spacing between conductors}}}$$

Figure 29: The factors that go into estimating the worst-case parasitic capacitance

Where possible, reduce the source $\frac{dv}{dt}$ or amplitude, though this often has functional tradeoffs.

And while we could next reduce the slew rate or voltage that the receiver is responsible for, that change often impacts the circuit functionally in an adverse way. The goal for mitigation of the impact is to either terminate or shunt the injected current away from the sensitive receiver.

This can result in:

1. In systems with cables and interconnects, this results in a properly terminated shield. This results in the noise currents being shunted to a common return through a low impedance connection
2. In PCB systems, guard traces, fencing, along with spacing to reduce and positively impact the parasitic capacitance is found, in addition to orthogonal routing.

These methods are shown in Figure 30.

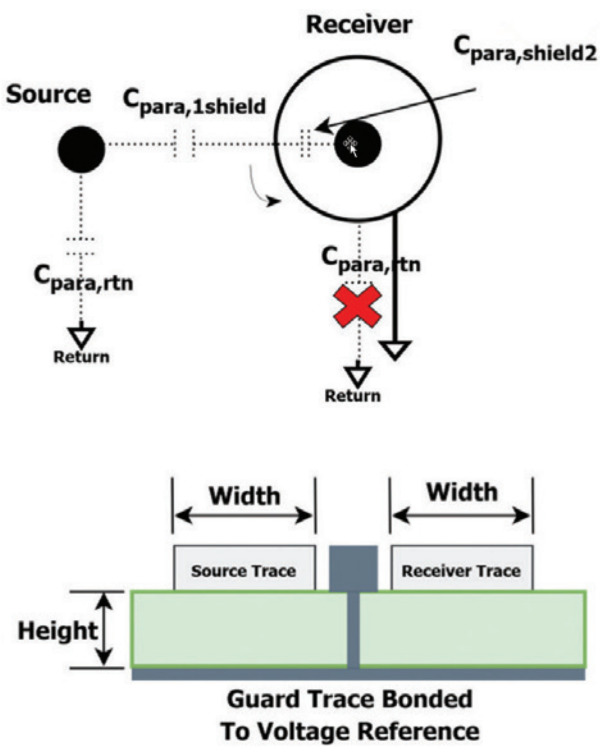


Figure 30: Managing parasitic capacitance in a cable (left), managing it with guard traces or via fencing (right)

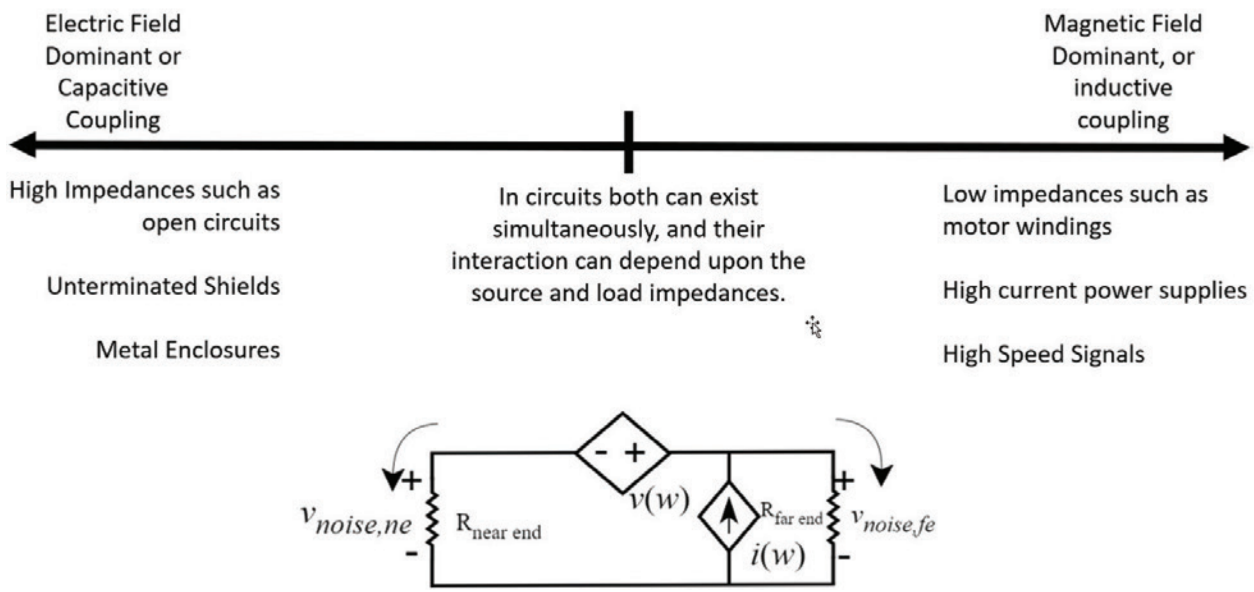


Figure 31: An example of the spectrum of near field coupling

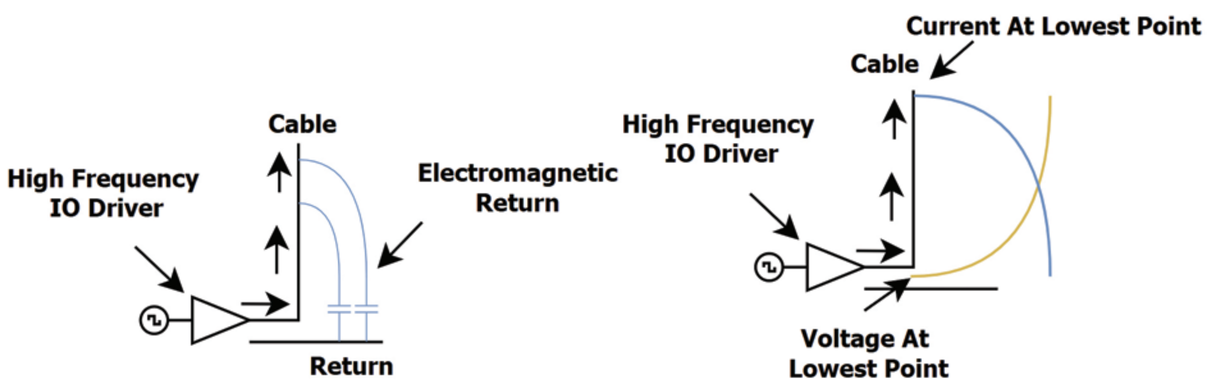


Figure 32: Current and voltage distribution in a cable (left), how cable drift current takes the electromagnetic return (right)

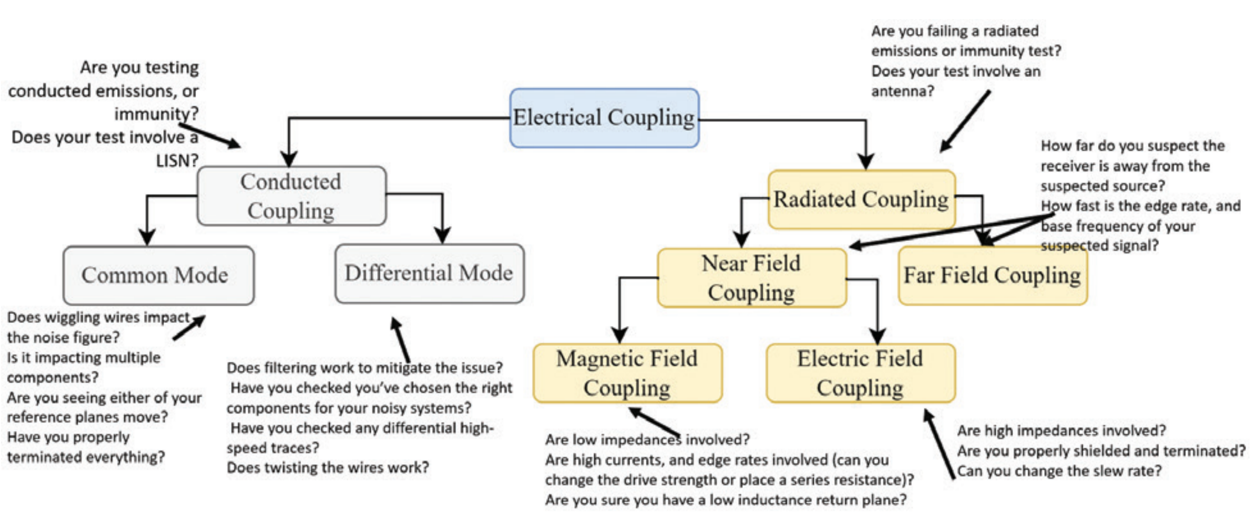


Figure 33: The answer to "you have a coupling problem"

In practice, magnetic and electric coupling do not exist apart from each other. The near field exists on a spectrum where loop geometry and frequency often result in problems with mixed signatures. As such, we can refer to Figure 31 to help organize troubleshooting efforts and determine what type of coupling your system is experiencing.

Far Field Coupling

Far field coupling is what you get when the receiver is far enough from the source that the fields behave like a traveling wave. The electric and magnetic parts are in phase, their ratio is about 377 ohms, and the wavefront is close to planar. In this region, long conductors behave like antennas. A cable or trace that carries common mode current looks like a monopole over the reference, with strong radiation near quarter wavelength and its odd multiples. Shorter lengths still radiate, just less efficiently. Field strength scales with common mode current and with the effective length of the conductor and falls with distance.

You can picture a driver launching current onto a cable shown in Figure 32.

The cable does not need a separate conduction return to radiate. Energy leaves as an electromagnetic wave, and the "loop" closes locally through displacement capacitance to the chassis and the environment near the source. That common mode current is what the chamber measures when the long cable acts as the antenna.

CONCLUSION

The next time you have a hard-to-decipher signal on your oscilloscope and a senior engineer says, "You have a coupling problem," pause before chasing symptoms. Think about how many paths let noise move from one part of a circuit to another.

Use the framework you have built and first identify the source, the path, and the receptor; decide whether the dominant mode is differential or common; and check whether a conducted issue has turned into an antenna. Figure 33 is not exhaustive, but it should keep you oriented and help you work through the next problem faster.

COEXISTENCE TESTING FOR WIRELESS MEDICAL DEVICES

Ensuring Reliable Performance in a Crowded Spectrum

My first exposure to wireless networking was when it became possible for me to send files to my printer over the 2.4 GHz band. That was almost 30 years ago. Data rates were not really fast, but fast enough to print documents or send e-mails. Since then, the number of devices using the unlicensed wireless spectrum and the available frequency band has grown significantly. Other than just being an interesting technology trend, many device manufacturers have found that using a wireless interface is more efficient and convenient than using traditional wired interfaces.

This is especially true for medical devices, especially in crowded emergency and operating rooms. (As a side note, I just had a doctor's visit this morning and noticed the number of devices that still had wired interfaces to their sensors.) Multiple equipment with multiple wired interfaces can get in the way of lifesaving measures and can even present trip hazards.

WIRELESS MEDICAL DEVICES IN HEALTHCARE SETTINGS: SOME BACKGROUND

It should not be a surprise, then, that medical devices have a good reason to shift from wired to wireless interfaces.

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By William Koerner

Figure 1 shows an AI-generated view of the growth of medical devices with wireless interfaces from 2020 to 2024.[1]

Here are some key insights from that study:

- Wearable devices lead the market, growing from $5.73B in 2020 to $8.85B in 2024, driven by fitness trackers, smartwatches, and wireless ECG monitors;
- Implantable devices (e.g., pacemakers, neurostimulators) show steady growth, reaching $6.27B in 2024;
- Handheld and portable devices (e.g., wireless glucose meters, ultrasound systems) are gaining traction due to mobility and ease of use.
- Stationary devices (e.g., wireless imaging systems) remain the smallest segment due to limited portability; and
- The "Others" category includes emerging technologies and niche applications, growing to $2.91B by 2024.

Today, with the increase of devices using wireless interfaces, there is an increased risk of interference. While the unlicensed frequency spectrum is "free," that does not mean that the use of that spectrum will always be available. Along with that, the use of the frequency spectrum is very busy. Figure 2 on page 32 shows the allocated frequency spectrum for the 2-3 GHz Band.

Medical devices will have to contend with transmitters that are either sharing the same channel, or even those that are not on the current operating channel but close enough in frequency to cause co-channel interference. On top of that, just about every patient and hospital employee will carry at least one wireless

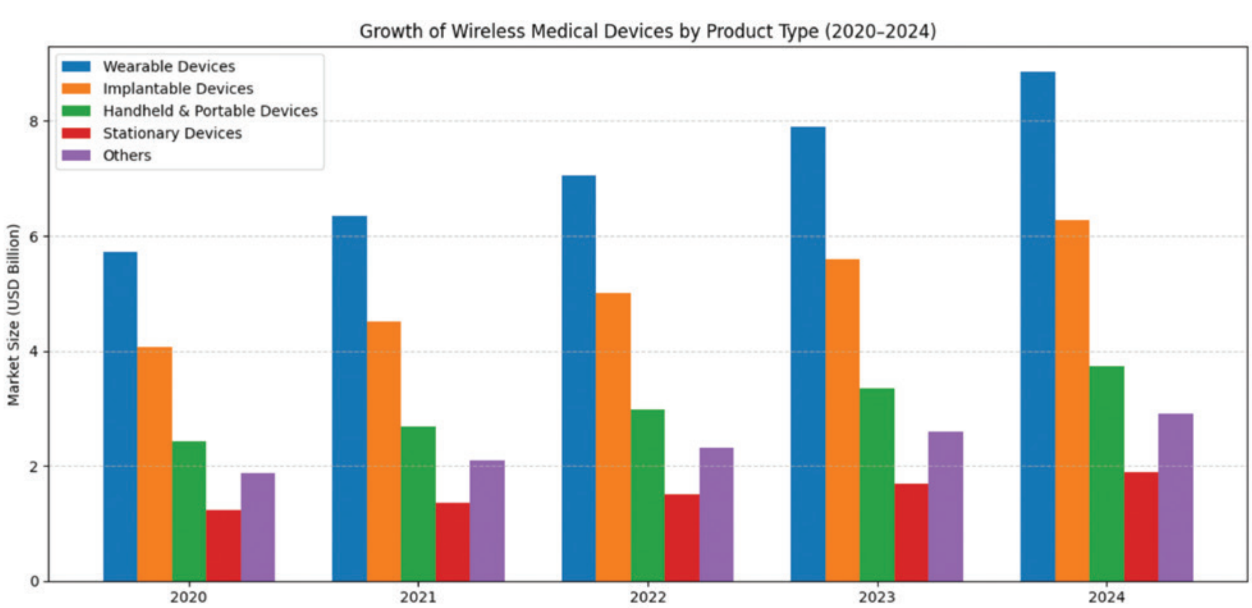


Figure 1: Growth of wireless medical devices by product type, 2020-2024

device with them, causing an even higher potential for interference.

In addition, visitors enter the hospital with smartphones configured to be wireless hotspots or connected to Bluetooth headsets, fitness bands, and smart watches. People change locations, make phone calls, and access Wi-Fi and LTE data networks and other wireless services during their visit, adding to the RF energy. At times, the “computers-on-wheels” and portable X-ray machines traveling the hallways periodically upload huge image files.

Yes, coexistence is a problem, not only in hospitals but anywhere large numbers of wireless devices are deployed. But wireless communication in a healthcare environment has become especially difficult. There are so many devices using RF that some are unable to establish or maintain a communications link at the level required for satisfactory operation. Devices that formerly worked in the engineer lab are failing to operate as required in the healthcare setting. The problem is further exacerbated by systems designed to support home healthcare technologies and that operate in home environments where the Internet of Things (IoT) includes radios to door locks, video doorbells, cameras, thermostats, and home appliances.

Ultimately, this leads to the risk of interference causing device malfunctions, data loss from patient monitoring devices, and potential impacts on patient safety. The purpose of this article is to explain the principles, challenges, and methodologies of coexistence testing from the perspective of a medical device manufacturer. Similar issues would exist for a hospital/clinic setting, but with a slightly different perspective (perhaps a future article?).

THE REGULATORY CONTEXT FOR COEXISTENCE TESTING

Definition of Coexistence Testing

You may be thinking this sounds like EMI or EMC, and you are only partly correct. Assessments for EMI and EMC issues primarily evaluate radio frequencies other than the wireless frequency used by medical devices. EMI and EMC test the purity of transmissions and sensitivity to high external fields other than RF channels in use by the equipment under test (EUT). Coexistence testing uses the same and close-by radio frequencies, and does the test using the desired EUT functionality, measuring effects of interference on the behavior of the EUT.

Relevant Standards and Guidance

In the U.S., the Federal Communications Commission (FCC) is responsible for defining the rules and

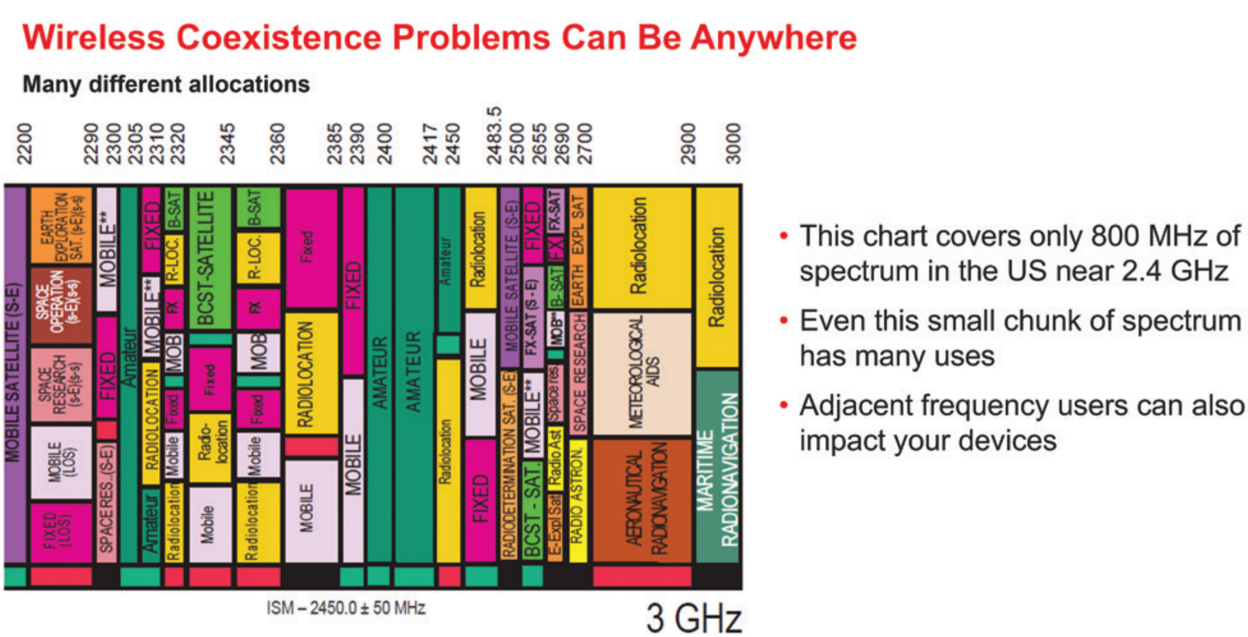


Figure 2: Frequency allocations for the 2.4 GHz band in the U.S.[2]

specifications for devices that use the frequency spectrum. The Food and Drug Administration (FDA) is responsible for establishing the requirements for medical devices and has recently had to address medical devices that integrate wireless interfaces. While the FDA does not define the rules and specifications for that wireless interface, it did have to start addressing issues caused by interference to those wireless interfaces.

Here are some of the relevant regulations, guidance documents, and standards applicable to coexistence testing of medical devices.

Radio Frequency Wireless Technology in Medical Devices - Guidance for Industry and Food and Drug Administration Staff[3]

This FDA-issued guidance, initially released in 2007 and updated in 2013, recommends coexistence testing for wireless devices intended for use in health care. From that document, the following requirements are indicated before filing for FDA approval.

From section 4b. of that document:

1. *Wireless Quality of Service – The submission should include information to describe the wireless QoS needed for the intended use and use environment of the medical device. This includes addressing any risks and potential performance issues that might be associated with data rates, latency, and communications reliability as described in Section 3-b.*
2. *Wireless coexistence – Any risks and potential performance issues that might be associated with wireless coexistence in a shared wireless environment should be addressed via testing and analysis with other wireless products or devices that can be expected to be located in the wireless medical device's intended use environment. See Section 3-c.*

Please note that the FDA considers this a guidance document, not a standards document, and also states:

> *This guidance represents the Food and Drug Administration's (FDA's) current thinking on this topic. It does not create or confer any rights for or on any person and does not operate to bind FDA or the public. You can use an alternative approach if the approach satisfies the requirements of the applicable statutes and regulations. If you want to discuss an alternative approach, contact the FDA staff responsible for implementing this guidance. If you cannot identify the appropriate FDA staff, call the appropriate number listed on the title page of this guidance.*

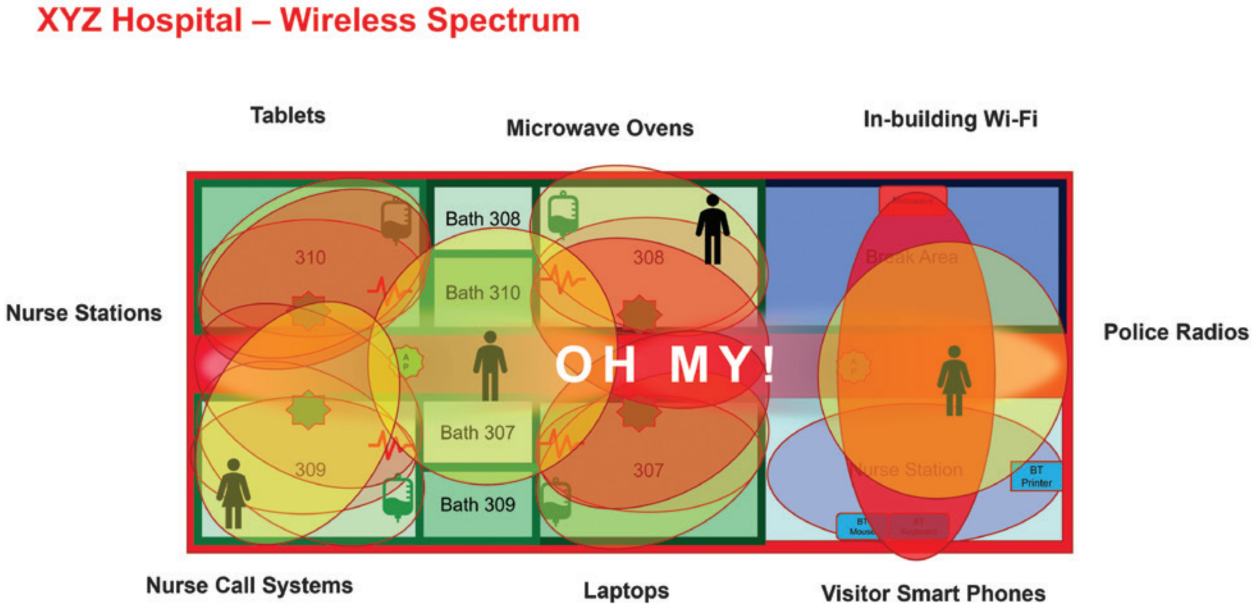


Figure 3: Example of RF crowding in hospital settings

ANSI C63.27-2021 - American National Standard for Evaluation of Wireless Coexistence[4]

Appendix A of the FDA Guidance Document lists reference documents to be reviewed for assessing the RF interfaces for medical devices. While that Appendix lists mostly EMC-related documents, the link in that section for all references does include a listing for ANSI C63.27, published in 2022 (after the publication of the FDA Guidance Document). This standard provides methods for evaluating the ability of a device to coexist in its intended RF wireless communications environment. While this is a recognized standard, it is by no means all that is required to meet the coexistence requirements of the FDA. View this standard as more of a minimum requirement (for reasons discussed later in this article).

AAMI TIR 69: Association for the Advancement of Medical Instrumentation - Risk Management of Radio-frequency Wireless Coexistence for Medical Devices and Systems (2017)[5]

This Technical Information Report (TIR) was developed by The Association for the Advancement of Medical Instrumentation (AAMI). It provides a process for defining risk management for medical devices that incorporate a wireless interface. It also references ANSI C63.27 for the recommended testing procedures, and the International Standards Organization (ISO) 14971 standard for risk management.

Regulatory Expectations

From the FDA guidance document, the following types of planning and execution are critical to a successful application for coexistence management:

- A summary of the coexistence testing, set-up, findings, and analysis;
- The wireless products (interferers) that were used in the coexistence testing, and their wireless RF frequencies, maximum output powers, and separation distances from the device;
- The specific pass/fail criteria for this testing;
- How the device and wireless functions were monitored during the testing and determined to meet the pass/fail criteria; and
- If it is reasonable to expect multiple units of the subject wireless medical device to be used in the same vicinity, the information should also address how the association and security between devices is established and maintained to prevent crosstalk among the devices.

The most challenging part of coexistence testing then is to determine the risk management plan to find the best fit for the specific EUT in its typical operating environments.

WIRELESS TECHNOLOGIES IN MEDICAL DEVICES

A key step in assessing the risk is to look at the wireless technologies that are implemented in the medical device. While the availability of different technologies and associated costs is very attractive, each has its own unique performance characteristics that will have a different response to coexistence with other wireless devices. Listed below are typical interface technologies for medical devices, and a description of their unique capabilities and performance.

Bluetooth®

Bluetooth comes in several different flavors, known as Bluetooth Classic and Bluetooth Low Energy (LE). Bluetooth Classic is the original implementation of Bluetooth Technology. It is a frequency hopping spread spectrum (FHSS) technology that uses a random hopping sequence of 79 possible channels with a 1 MHz bandwidth from 2402 to 2480 MHz. Because it uses a frequency hopping technique, it tends to handle interference from other sources better than other wireless technologies. Versions 1-3 of Bluetooth Classic have a maximum range of up to 10 meters.[6]

Bluetooth Low Energy (BTLE) is an extension to the original Bluetooth and was designed specifically to allow for lower energy use and longer battery life. Like Bluetooth Classic, it uses multiple channels, but in a different scheme. Three channels are dedicated as advertising channels and are not used for data transmission. It also utilizes a 1 MHz bandwidth in the same frequency range, with channels spaced every two MHz. Because of the wider channel spacing, there are less (40) total channels available. With half of the channels available for transmission, it can make BTLE more susceptible to interference compared to Bluetooth Classic.

Versions 4 and 5 of Bluetooth have a maximum range of up to 60 meters, and possibly 240 meters for BTLE Long Range.[7] Because of the ease of implementation of either type of Bluetooth, this technology can be found in medical devices, especially those that are partnered with an application that runs on a smartphone or PC.

Zigbee[8]

Zigbee is an IEEE 802.15.4-based specification for a suite of high-level communication protocols used to create personal area networks with small, low-power digital radios, such as for home automation, medical device data collection, and other low-power, low-bandwidth needs for small-scale projects that need wireless connection. It uses a 2 MHz bandwidth, spaced every 5 MHz, in the 2.4 GHz band with a maximum of 16 channels. Zigbee has a maximum operating range (non-mesh) of 10-20 meters. Because this protocol uses the 2.4 GHz bands, it has all the interference considerations of other standards addressing this frequency band.

Wi-Fi

Wi-Fi is probably the largest wireless technology used for unlicensed spectrum. At least ten (10) different Wi-Fi standards have been released over the last 25 years that may be applicable to medical devices.[9] This technology relies on an access point and paired station (or client) devices to transmit data back and forth. Wi-Fi also lends itself to providing an Internet connection for the Station device, allowing control and requesting information from anywhere in the world.

Here is a brief description of the different Wi-Fi standards, with an emphasis on their use in the U.S.:

- *802.11b:* Although it is the second Wi-Fi standard developed by the IEEE, it was the first implemented in September 1999. It uses the 2.4 GHz band, specifically 2401-2473 MHz. The channel bandwidths are 22 MHz, and it allows for 11 channels in the U.S. However, that does not mean that all 11 channels can be used in the same area, as 802.11b has overlapping channels. This standard has a maximum range of up to 35 meters.
- *802.11a:* Released in September 1999, this standard makes use of the 5 GHz band only. It was the first expansion into the newly available 5 GHz band, allowing for higher data rates, but still limited to a 20 MHz bandwidth maximum. This was not an overlapping technology, thus allowing for up to 31 available channels and a maximum operating range of 35 meters.
- *802.11g:* Released in June 2003, this standard uses only the 2.4 GHz band and a maximum bandwidth of 20 MHz. The smaller bandwidth allows for up to 4 channels in the band without overlapping. It has a maximum operating range of 38 meters.
- *802.11n:* Released in October 2009, this was the first Wi-Fi standard to be branded by the Wi-Fi Alliance, known as Wi-Fi 4. It allows for both 20 and 40 MHz bandwidths and uses both the 2.4 and 5 GHz bands. Because of this, it also allows for a larger number of available channels, up to 43 for 20 MHz bandwidths, and 23 40 MHz bandwidths. It has a maximum operating range of up to 70 meters.
- *802.11ac:* Released in December 2013, this standard is also known as Wi-Fi 5. It uses the 5 GHz band only, but allows for 20, 40, 80, and 160 MHz bandwidths, allowing for even higher data rates. It allows for 31-20 MHz, 14-40 MHz, 7-80 MHz, and 3-160 MHz channels. It has a maximum operating range of up to 30 meters.

- *802.11ax:* Released in May 2021, this standard is also known as Wi-Fi 6 (using 2.4 and 5 GHz bands only) and Wi-Fi 6E (E=Extended), which uses all three bands. This standard truly increased the capacity and capability of Wi-Fi, allowing for up to 103-20 MHz, 52-40 MHz, 21-80 MHz, and 10-160 MHz channels. It has a maximum operating range of up to 30 meters.
- *802.11be:* Released in September of 2024, this standard is also known as Wi-Fi 7 and uses all three bands. It allows for even wider bandwidths, adding 240 and 320 MHz channels. It also introduces the mandatory use of pre-amble puncturing, a technique that allows the device to notch out part of the channel in the presence of interference. This makes it one of the first standards to allow for managing interference in the operating channel. (Actually, this capability was available with Wi-Fi 6E but was optional). This standard has a maximum operation range of up to 30 meters.
- *60 GHz Wi-Fi:* There are three different Wi-Fi standards using the 60 GHz band (802.11ad, aj, and ay) that are defined as multi-gigabit standards. They offer data rates between 1.08 GHz and 8.64 GHz with a maximum operating range up to 10 meters for 802.11ay. Because of the data rates, this standard may be used for applications that require very high data rates, such as high-resolution digital video. This band, while limited in operating range, would probably be the most immune to interference, as there are currently not many devices transmitting in this band.

A Word about Frequency Bands

As described in this section, Wi-Fi can use potentially up to three different frequency bands. Each has its benefits and limitations, as detailed here:

- *2.4 GHz:* 2.4 GHz allows for longer transmitting distances but is limited in the number of channels and has to deal with multiple other technologies using this band. Microwave ovens operate in this band and can represent a significant interference source as well.
- *5 GHz:* The 5 GHz bands allow for more channels and wider bandwidths (up to 240 MHz) and are popular for devices that use wider bandwidths and higher data rates. However, the 5 GHz band has some incumbent users that have priority access to the bandwidth. These include various forms or radars using the 5250-5350 MHz, 5470-5730 MHz bands. For that reason, APs and some station devices that use these frequencies must employ a radar detection function that monitors for defined radar signals and, if such signals are detected, must stop transmitting in that channel and not attempt to re-use that channel for up to 30 minutes (this is known as dynamic frequency selectivity, or DFS). It is possible to design the interface to avoid those bands, but that reduces the number of available channels to eight 20 MHz channels.
- *6 GHz:* The 6 GHz band was opened for unlicensed devices in the U.S. in April 2020. The U.S. was the first country to allow use of the entire 6 GHz band, covering from 5925-7125 MHz. This allows for many channel possibilities, but, like the 5 GHz band, comes with some restrictions. There are multiple licensed (incumbent) users in the 6 GHz spectrum. As part of the agreement to use this band, any unlicensed device that uses this band must employ a contention-based protocol (CBP) in the U.S. Similar to the DFS requirements, devices using this band must also employ a receiver detection function that monitors for incumbent transmitters, and if detected above a threshold value, cease transmissions in this channel. Unlike the 5 GHz band, the device does not need to move from this channel, but can wait for the incumbent to stop transmitting, and then reuse the channel.

Cellular

In the past few years, device manufacturers have started integrating cellular wireless technologies into their devices. These technologies can use frequencies anywhere from the 700-900 MHz range, the lower and upper parts of the 2 GHz band, and most of the 5-6 GHz bands. These present potential co-channel interference issues with devices using the 2.4 GHz bands, and even the 5 and 6 GHz bands. Because these frequencies are licensed to use the assigned spectrum, they have priority over unlicensed devices operating in the same spectrum.

COEXISTENCE TESTING METHODOLOGIES

As mentioned previously, the first step is to develop the risk plan for your product and, from that, develop a comprehensive test plan rooted in the risk-based approach and using the standards listed previously.

This includes:

Defining the Intended Use Environment

Is the device something that would only be used in an operating room? Or is it a wearable device that could be used anywhere? Will it be located in a static environment (fixed location) or mobile (such as in an ambulance)? This is the starting point to determine just how big or complicated a testing scenario you will need to develop.

Identify Wireless Technologies and Co-Located Systems

Secondly, build a list of wireless technologies used in your device, and the potential co-located systems operating in the same band or adjacent bands. Consider all potential medical (other medical devices with wireless interfaces) or non-medical (e.g., hospital APs, cellular phones, patient and visitor Bluetooth devices and locations), etc.

Conduct Spectrum Analysis

Ideally, as a device manufacturer, it would provide great insight if there were an actual spectrum analysis or signal monitoring of the intended use environment (this is the part I mentioned earlier as applying more to the hospital or clinic environment). Having an actual measured expected spectrum makes it easier to develop a targeted interference test plan.

Risk Assessment Matrix

With the previous information, evaluate the likelihood and severity of the interference-related failures that could happen. This is where the functional wireless performance (FWP), whether or not the device can perform its intended function and performance) comes into play. Be sure to include the following potential risk areas:

- *Criticality of device function:* Does this device provide life-sustaining functionality, or just monitoring?
- *Impact of communication failure:* Can the device still meet its functional performance with a loss of data or delayed alerts?
- *Recovery mechanisms:* Is there a designed recovery mechanism? Can the device recover by re-transmitting the lost or corrupted data? Is there a fallback, such as an audible or visible alert on the device?

Define Performance Metrics

Once you have decided which risks need to be tested, you then need to consider how to measure the various performance metrics to determine if your device meets the FWP requirements. This includes establishing quantitative thresholds for acceptable performance under interference, such as:

- Packet error rate (PER)
- Latency
- Throughput
- Connection stability
- Sensitivity testing
- Time to recover

Test Set-up and Execution

The ANSI C63.27 standard defines three tiers to coexistence testing. Each tier is defined for different intended performance requirements for the medical device, as follows;

- Tier 1 represents the most thorough type of evaluation. Its purpose is to test those devices that have the highest consequences of unacceptable performance, or where the highest levels of uncertainty are required. This includes a wider range of unintended interference signals and the potential for interference from adjacent channel interference. This tier is for those products whose functional performance is the most critical.

- Tier 2 represents the mid-level type of evaluation. This Tier reduces the number of interference signals from Tier 1, and some testing is done for potential interference from adjacent channel interference. This Tier is for products that do not have such a critical performance requirement are not considered life-sustaining devices.
- Tier 3 represents the lowest level of interference testing. The intention is to provide the greatest insight into the EUT coexistence capabilities with the most limited amount of testing.

Annex A of the standard provides guidance for the types of interference scenarios that should be used based on the wireless interference in the EUT. For example, if the EUT employs just a Bluetooth interface, the recommended test signals include a combination of just a IEEE 802.11n Wi-Fi signal for Tier 3 to multiple IEEE 802.11n Wi-Fi signals and several adjacent band LTE signals for Tier 1.

Figure 4 shows an example of testing a Bluetooth LE device for all three Tier 1 scenarios.

However, please keep in mind that this does not represent all that is required for coexistence testing but should be viewed as a minimum set of test scenarios. For example, if you know that your device would be considered a mobile device, you may want to consider adding scenarios where the interference signals change level with time to simulate moving towards or away from another wireless device. You may also want to consider the sensitivity of your EUT and vary the levels of the interference signal until the device fails to meet its KPIs.

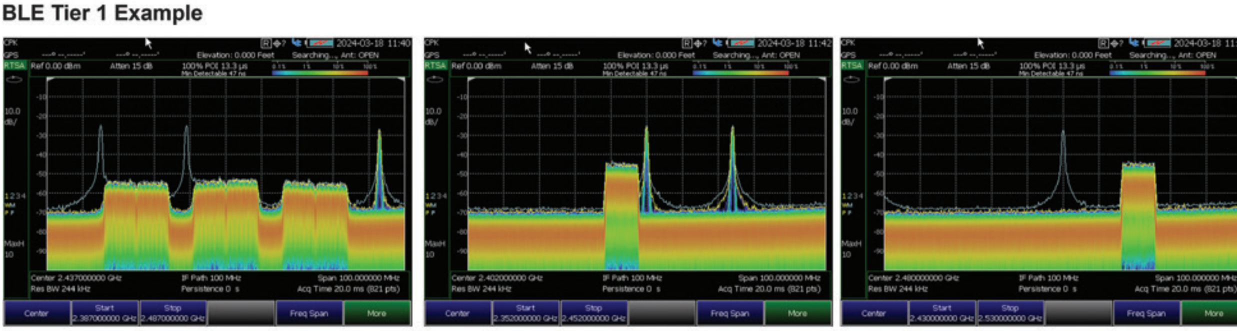
TEST ENVIRONMENTS

The ANSI C63.27 standard allows for four different testing environments, as follows:

1. *Conducted:* This environment is a fully conducted testing environment. The antennas are removed from the EUT, and all interference signals are fed into the EUT or companion device through cables. This environment allows for the most controlled testing, as you will not have to consider in-building Wi-Fi or Bluetooth signals being included in your testing. It also allows for the ability to use external components, such as variable attenuators, splitters, etc. for maximum control of the interference to EUT signal ratios.
2. *Dual Chamber:* This environment moves towards a more realistic testing environment, as the EUT antennas are included in the test, and both the EUT and companion are in a shield chamber, virtually eliminating any external interference signals. All interference signals are then fed through antennas in each chamber. This environment is a bit challenging in that setting a desired interference signal level into the EUT takes some calculation of external measurements. Your test solution should provide a way to easily determine the radiated interference signal level to the EUT.
3. *Full Anechoic:* This environment allows for the most control in testing the EUT and offers the most controlled testing environment. It shares similar issues to the dual chamber environment in determining the interference signal levels into

C63.27 Interference Signals

BLE Tier 1 Example



C63.27 A.2.2.1 (BLE Tier 1 Set 1) | C63.27 A.2.2.2 (BLE Tier 1 Set 2) | C63.27 A.2.2.2 (BLE Tier 1 Set 2)

Figure 4: Example of Tier 1 testing for Bluetooth LE EUT

the EUT. It also requires an anechoic chamber, which, if you do not have one, is a very expensive investment.

4. *Over-the-Air (OTA):* This environment could be considered a poor man's chamber environment. You basically test over the air in an open space, such as an unused conference room. There might be challenges in getting the distances far enough for far field, not to mention having to anticipate the in-building Wi-Fi and Bluetooth that will interfere with your interference testing.

Table 1 provides a summary of the pros and cons of each testing method.

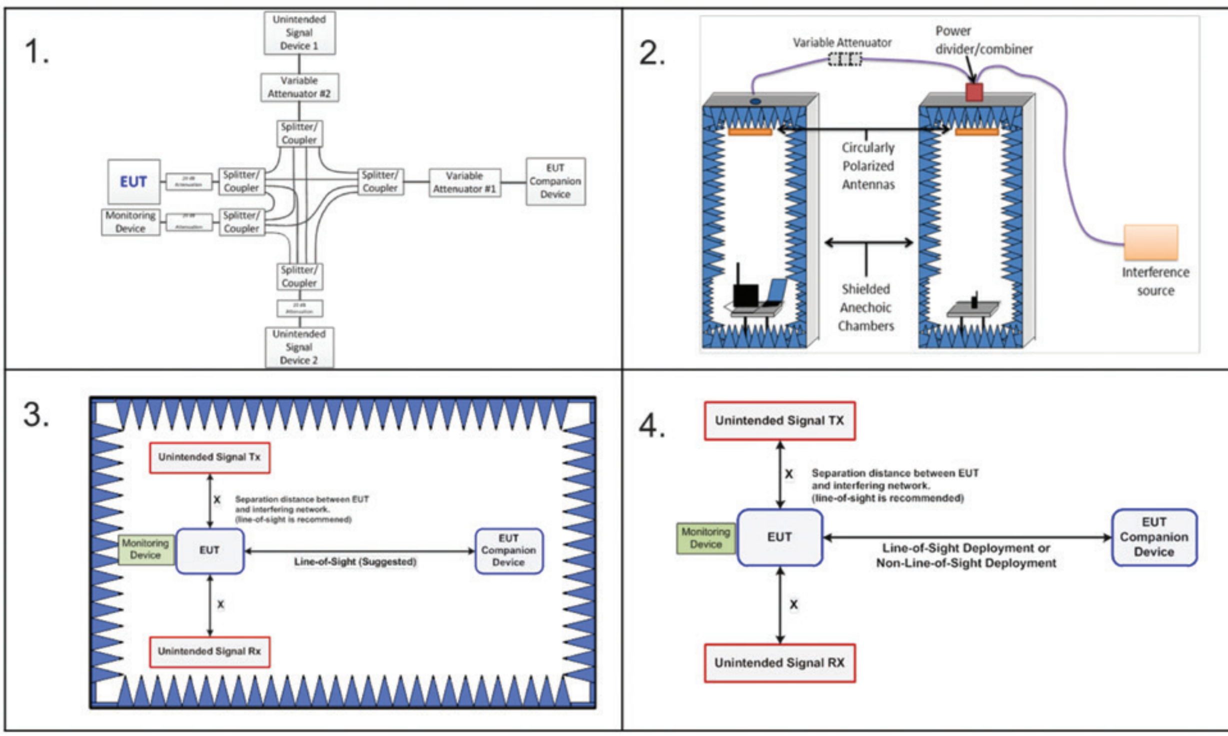


Figure 5: Different allowed testing environments for coexistence testing

Method	Pros	Cons	Best Use Case
Conducted	• High repeatability and control • Low cost • No RF interference	• Doesn't reflect real-world RF environment • Limited spatial effects	Early-stage testing, protocol-level coexistence
Two-Chamber	• Isolated environments • Better control of RF paths • Moderate cost	• Limited spatial realism • Setup complexity	Controlled RF testing with some spatial separation
Open Air	• Realistic RF environment • Easy setup • Good for field testing	• Low repeatability • Susceptible to external interference	Real-world performance validation
Anechoic Chamber	• Highly controlled RF environment • Eliminates reflections • High accuracy	• Expensive • Complex setup • Limited availability	Final validation, certification, high-precision testing

Table 1: Testing method, pros and cons

TEST SOLUTION

As you can see, a test solution can be as simple as a few commercially available Wi-Fi and Bluetooth devices. But, if you wish to extensively test beyond the requirements of ANSI C63.27 and consider roaming interferers, sensitivity testing, multiple variable interference sources, or perhaps a way to measure time to recover, it will require more than just commercial devices, and may still require a lot of manual analysis to fully evaluate your device.

Figure 6 shows an example of automated signal generators and signal conditioning to run complex interference testing. It also shows using software to read the display of the EUT application to determine if the KPI has met the desired functional performance. In this case, the software is monitoring the portion of the smartphone application to determine if the device is still connected to its companion device.

SUMMARY

While the idea of managing coexistence for medical devices has been around for over 15 years, the practice of what makes the best or correct test continues to be a challenge today. Fortunately, the test equipment and software have evolved to allow for the design of very complex test scenarios to verify that both the FWP and KPIs for the EUT are met. Doing this supports the design validation before costly changes need to be made to the design and helps to avoid multiple review rounds for FDA approval.

ENDNOTES

1. “Wireless Medical Devices Market Size | CAGR Of 12.1%”, Published: March 2025. Report ID: 141096
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7. ibid.
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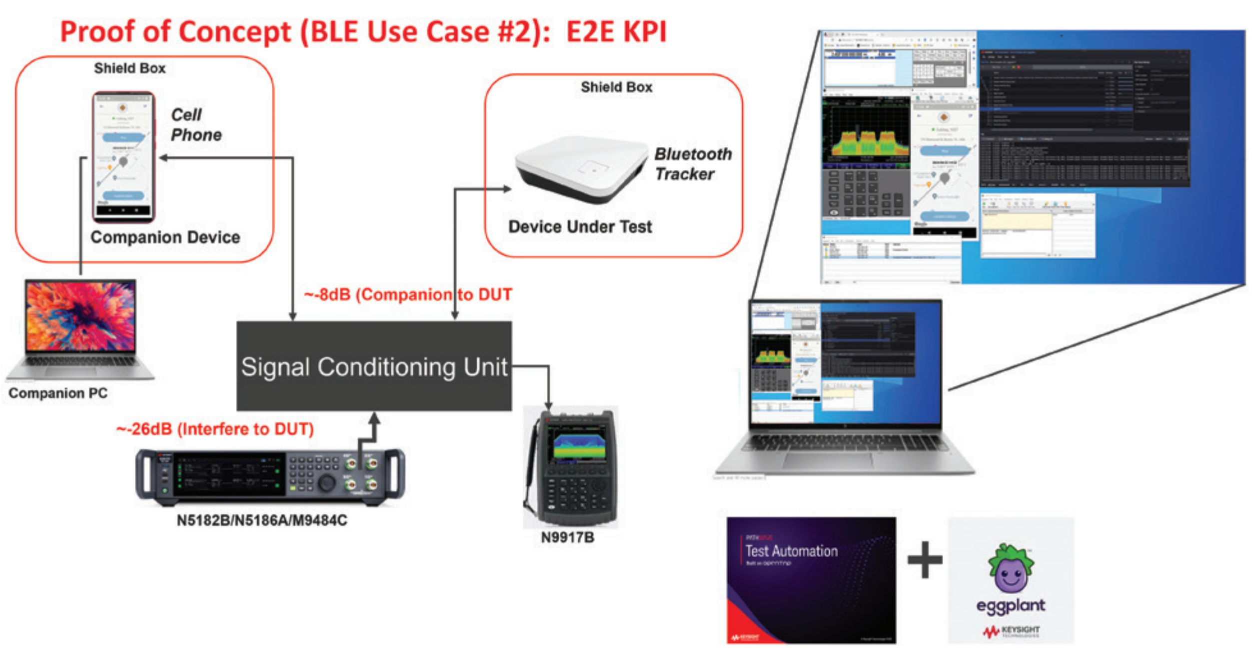

Figure 6: Automated coexistence testing solution with automated EUT application monitoring

SHIELDING TO PREVENT RADIATION

Part 6: Near-Field Shielding Effectiveness of a Solid Conducting Shield

By Bogdan Adamczyk

This is the sixth of seven articles devoted to the topic of shielding to prevent electromagnetic wave radiation. The first article [1] discussed the reflection and transmission of uniform plane waves at a normal boundary. The second article [2] addressed the normal incidence of a uniform plane wave on a solid conducting shield with no apertures. The third article [3] presented the exact solution for the shielding effectiveness of a solid conducting shield. The fourth article [4] presented the approximate solution obtained from the exact solution. The fifth article [5] discussed the wave impedance of electric and magnetic dipoles.
In this article, we will use the concept of wave impedance to determine the shielding effectiveness in the near field.

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NEAR-FIELD SHIELDING – ELECTRIC SOURCES

Note: The following derivations are valid under the assumption that the shield made of a good conductor is much thicker than the skin depth, at the frequency of interest.

The shielding effectiveness in the near field for electric sources is:

$$S_{dB,e} = R_{dB,e} + A_{dB} \tag{1}$$

The absorption loss in the near field is the same for the electric sources and is the same as it was in the far field [4]. That is:

$$A_{dB} = 20\log_{10} e^{\frac{t}{\delta}} \tag{2a}$$

or

$$A_{dB} = 131.434 \times t\sqrt{f\mu_r\sigma_r} \tag{2b}$$

when the conductor thickness is expressed in meters, or

$$A_{dB} = 3.338 \times t\sqrt{f\mu_r\sigma_r} \tag{2c}$$

when the conductor thickness is expressed in inches.

Near-field shielding formulas for the reflection loss can be derived using the far-field shielding results for the reflection loss and the concept of the *near-field wave impedance* discussed in the previous article [5].

The reflection loss of a good, thick conductor in the far field was derived in [4] as:

$$R_{dB} = 20\log_{10}\left|\frac{\eta_0}{4\hat{\eta}}\right| \tag{3}$$

The near-field wave impedance for electric sources was derived in [5] as:

$$\left|\hat{Z}_{w,e}\right| \cong 60\frac{\lambda_0}{r} \tag{4}$$

The reflection loss for the near-field electric sources, $R^{dB,e}$, is obtained by substituting this wave impedance for the intrinsic impedance of free space in Equation (3).

Thus,

$$R_{dB,e} \cong 20\log_{10}\left|\frac{\hat{Z}_{w,e}}{4\hat{\eta}}\right| \tag{5}$$

or

$$R_{dB,e} \cong 20\log_{10}\left|\frac{60\frac{\lambda_0}{r}}{4\hat{\eta}}\right| \tag{6}$$

Wavelength in free space can be expressed as:

$$\lambda_0 = \frac{v_0}{f} = \frac{3\times10^8}{f} \tag{7}$$

Now, recall Equation (26) from [4] for the magnitude of the intrinsic impedance of the shield:

$$\left|\hat{\eta}\right| = \sqrt{\frac{8\pi^2 f\mu_r}{\sigma_r(5.8\times10^{14})}} \tag{8}$$

Substituting Equations (7) and (8) into Equation (6), we get:

$$R_{dB,e} \cong 20\,log_{10}\frac{\frac{180\times10^8}{fr}}{4\times\sqrt{\frac{8\pi^2 f\mu_r}{\sigma_r(5.8\times10^{14})}}} \tag{9}$$

or [6]:

$$R_{dB,e} \cong 321.72 + 10\,log_{10}\frac{\sigma_r}{f^3r^2\mu_r} \tag{10}$$

Thus, the near-field shielding effectiveness for electric field sources is

$$S_{dB,e} \cong 321.72 + 10\,log_{10}\frac{\sigma_r}{f^3r^2\mu_r} + A_{dB} \tag{11}$$

Where A_{dB} is given by Eqs. (2).

NEAR-FIELD SHIELDING – MAGNETIC SOURCES

The reflection loss of a good, thick conductor in the far field was derived in [4] as:

$$R_{dB} = 20\log_{10}\left|\frac{\eta_0}{4\hat{\eta}}\right| \tag{12}$$

The near-field wave impedance for magnetic sources was derived in [5] as:

$$\left|\hat{Z}_{w,m}\right| \cong 2369\frac{r}{\lambda_0} \tag{13}$$

The reflection loss for the near-field magnetic sources, $R_{dB,e}$, is obtained by substituting this wave impedance for the intrinsic impedance of free space in Equation (12).

Thus:

$$R_{dB,m} \cong 20\log_{10}\left|\frac{\hat{Z}_{w,m}}{4\hat{\eta}}\right| \tag{14}$$

or

$$R_{dB,m} \cong 20\log_{10}\left|\frac{2369\frac{r}{\lambda_0}}{4\hat{\eta}}\right| \tag{15}$$

where

$$\lambda_0 = \frac{v_0}{f} = \frac{3\times10^8}{f} \tag{16}$$

$$\left|\hat{\eta}\right| = \sqrt{\frac{8\pi^2 f\mu_r}{\sigma_r(5.8\times10^{14})}} \tag{17}$$

Thus:

$$R_{dB,m} \cong 20\,log_{10}\frac{2369\frac{fr}{3\times10^8}}{4\times\sqrt{\frac{8\pi^2 f\mu_r}{\sigma_r(5.8\times10^{14})}}} \tag{18}$$

or [6]:

$$R_{dB,m} \cong 14.57 + 10\,log_{10}\frac{fr^2\sigma_r}{\mu_r} \tag{19}$$

Thus, the near-field shielding effectiveness for magnetic field sources is:

$$S_{dB,m} = 14.57 + 10\,log_{10}\frac{fr^2\sigma_r}{\mu_r} + 20\log_{10}e^{\frac{t}{\delta}} + A_{dB} \tag{20}$$

where A_{dB} is given by Eqs. (2).

NEAR-FIELD SHIELDING EFFECTIVENESS – COPPER VS. STEEL - SIMULATIONS

In this section, we compare the near-field shielding effectiveness of copper and steel (SAE1045). Table 1 shows the relative conductivity and relative permeability of these two shield materials.

Material	σ_r	μ_r
Copper	1	1
Steel (SAE 1045)	0.1	1000

Table 1: Relative conductivity and permeability of copper and steel

Let's begin with the reflection loss for electric field sources, at a distance of 5 mm, computed from Eq. (10), repeated here:

$$R_{dB,e} \cong 321.72 + 10\,log_{10}\frac{\sigma_r}{f^3 r^2 \mu_r} \qquad (21)$$

Figure 1 shows the electric field reflection loss in the frequency range 100 Hz – 1 GHz. Note that the reflection loss of copper is higher over the entire frequency range.

Next, we compare the reflection loss for magnetic field sources. It is calculated at a distance of 5 mm from the source and is computed from Eq. (19), repeated here:

$$R_{dB,m} \cong 14.57 + 10\,log_{10}\frac{f r^2 \sigma_r}{\mu_r} \qquad (22)$$

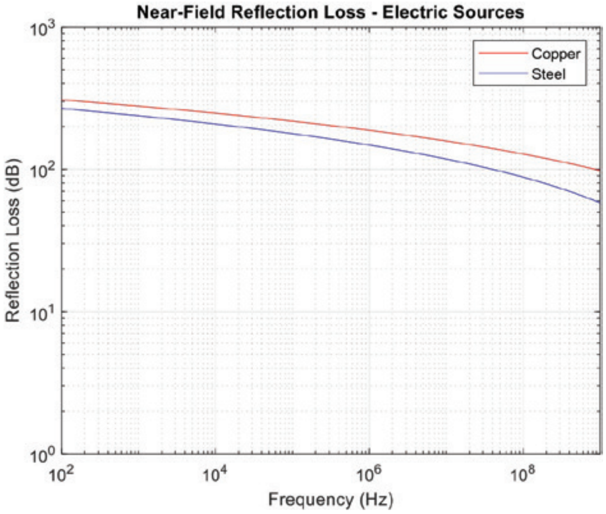

Figure 1: Reflection loss – electric field– copper vs. steel

Figure 2 shows the magnetic field reflection loss in the frequency range 100 Hz – 1 GHz.. Note that the reflection loss of copper is higher over the entire frequency range.

The absorption loss, for 20-mil thick shields, is calculated from Eq. (2c), repeated here:

$$A_{dB} = 3.338 \times t\sqrt{f \mu_r \sigma_r} \qquad (23)$$

and is shown in Figure 3. Note that the absorption loss of steel is higher over the entire frequency range.

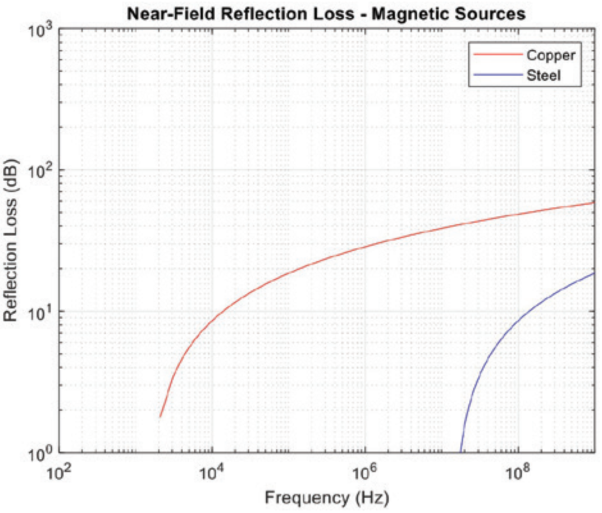

Figure 2: Reflection loss – magnetic field– copper vs. steel

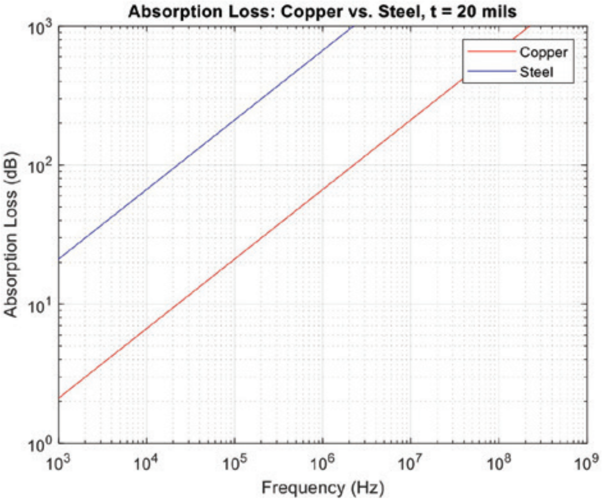

Figure 3: Absorption loss – copper vs. steel

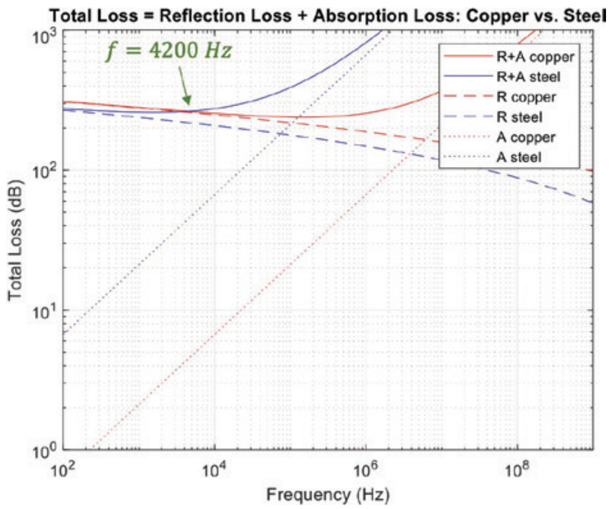

Figure 4: Total shielding effectiveness – electric field sources - copper vs. steel

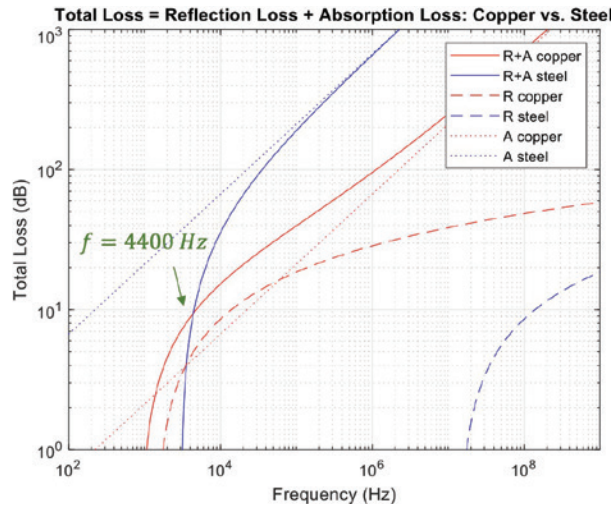

Figure 5: Total shielding effectiveness – magnetic field sources - copper vs. steel

The total shielding effectiveness for electric field sources, shown in Figure 4, is calculated from:

$$S_{dB,e} = R_{dB,e} + A_{dB} \tag{24}$$

where and are calculated from Eq. (21) and (23), respectively.

Note that up to the frequency of about 4200 Hz, the shielding effectiveness of copper is higher than that of steel. Beyond that frequency, the opposite is true.

The total shielding effectiveness for magnetic field sources, shown in Figure 5, is calculated from:

$$S_{dB,m} = R_{dB,m} + A_{dB} \tag{25}$$

where $R_{dB,e}$ and A_{dB} are calculated from Eq. (22) and (23), respectively.

Note that up to the frequency of about 4400 Hz, the shielding effectiveness of copper is higher than that of steel. Beyond that frequency, the opposite is true.

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CAN MECHANICAL MOVEMENTS ON FI-CDM TESTER CAUSE ADDITIONAL ZAP DURING CDM STRESS?

By Sheela Verwoerd for EOS/ESD Association, Inc.

This investigation is the first to examine how mechanical movements impact CDM stress, revealing that such movements add stress to the Device Under Test—a finding relevant for Field-Induced CDM testers and any tester requiring part movement to trigger a zap. The study provided clear evidence of unwanted zaps caused by contact bouncing and proposed mitigation solutions. Based on these findings, the tester supplier modified the tester's software, representing evolutionary but significant progress in improving test accuracy and reliability.

During FI-CDM testing, discharge current polarity can sometimes oppose the stress condition. In Figure 1, a negative zap occurs despite a +500V stress condition. We refer to this unintended zap as a secondary discharge.

Secondary discharges occur inconsistently and only in certain package types, specifically in Single Discharge mode.

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During CDM stress, the test part is placed in a dead-bug position on a field plate. A CDM discharge is triggered when the pogo pin contacts the part [1]. Both the discharge head and pogo pin are grounded. The discharge head's movement is motor-controlled, guided by software and the x, y, z coordinates set during part alignment [2].

IS SECONDARY DISCHARGE POSSIBLE IN THEORY?

Let's examine the tester architecture and operation to determine if mechanical vibrations on the pogo pin can cause secondary discharge. Figure 2 illustrates the sequence of events during a single zap in SD mode.

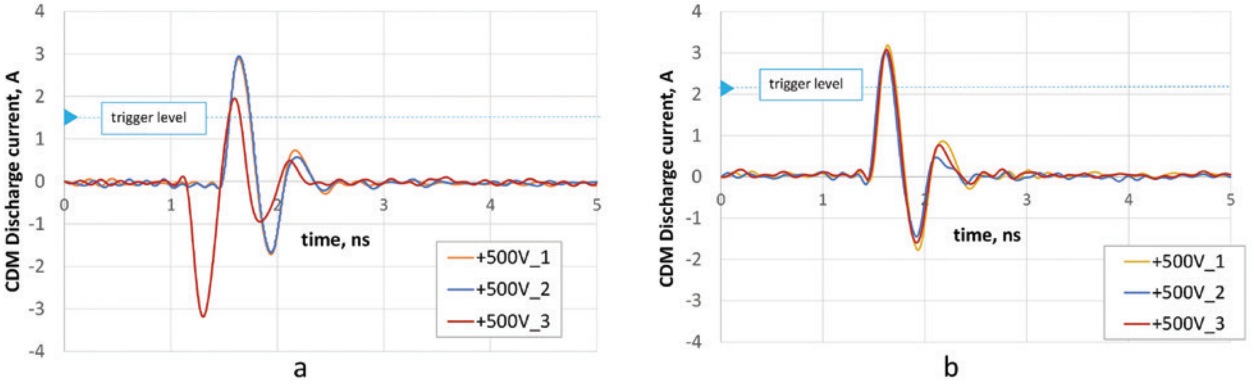

Figure 1: CDM waveform data at TC+500V on H8 with different trigger levels.

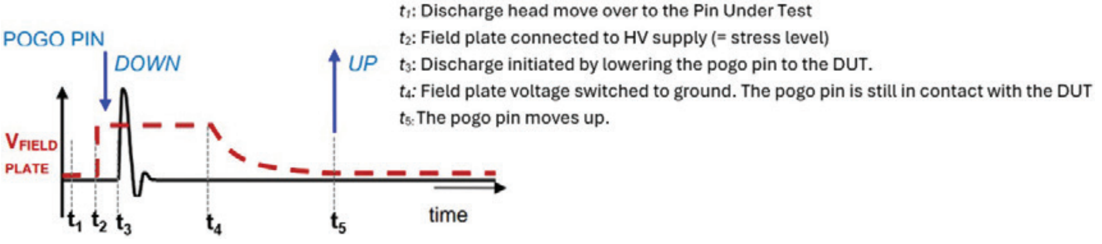

Figure 2: Single Discharge procedure (Annex H of JS-002)

Figure 3 explores the charge and voltage on the IC during the different steps mentioned above.

Steps 1 to 3 proceed as expected. However, if the pogo pin loses contact in the down position due to misalignment or mechanical bouncing (Step 4), a subsequent contact may trigger a secondary discharge. This occurs only if the field plate voltage differs from that at the previous contact in Figure 4.

PROOF OF SECONDARY DISCHARGE

Now that we have established its theoretical possibility, it must be validated through measurements. To study this event, we first need to make it reproducible. This was achieved by applying CDM stress to a verification coin, deliberately misaligning the pogo pin to touch its edge and ensure contact loss.

Measurement 1: Monitor field plate voltage and CDM discharge currents simultaneously

Figure 4 illustrates the measurement setup used for this study, with standard equipment hardware shown in blue and adaptations in red. A deep-memory oscilloscope (Scope-2) with a high sampling rate and 1 GHz bandwidth monitors both the CDM current and the field plate voltage simultaneously, as shown in Figure 5.

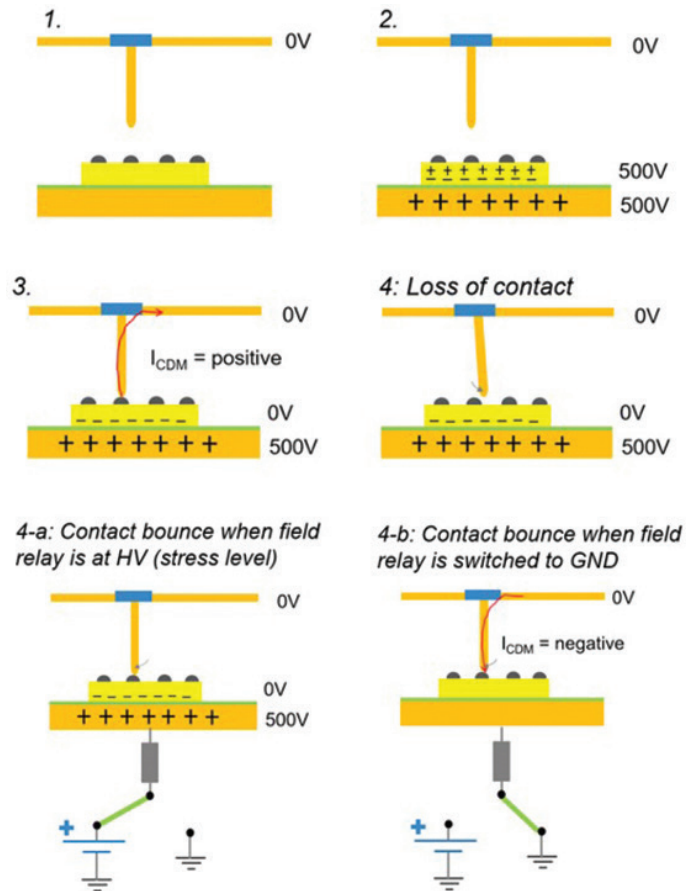

Figure 3: Charge and voltage values during a CDM event at TC+500 in SD mode

From the several measurements, we see the following:

a. In the SD mode, secondary discharge occurs just after the voltage on field plate switches from HV to ground.

b. The discharge time set in the test program does not have any impact on the duration of the HV on field plate nor the time at which the secondary discharge occurs.

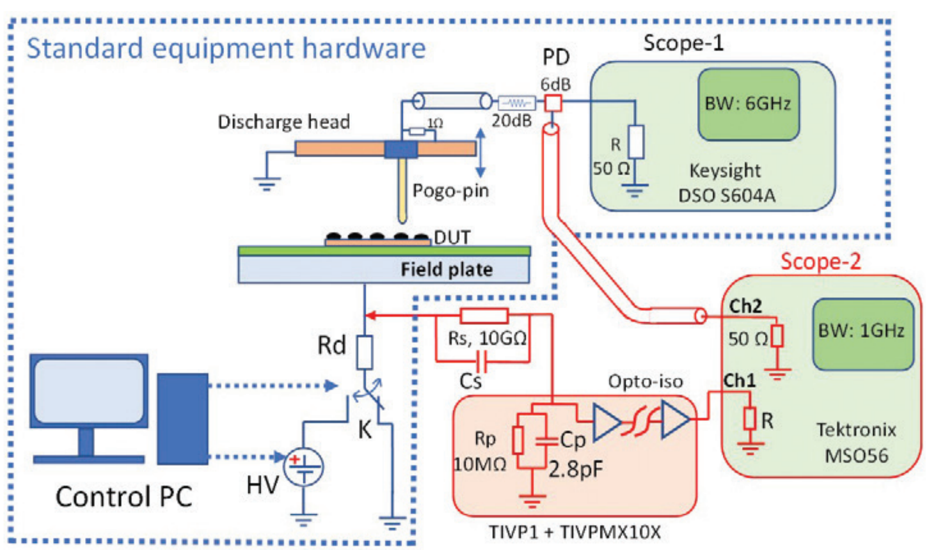

Figure 4: Measurement setup for simultaneous monitoring of CDM pulse and charge plate potential.

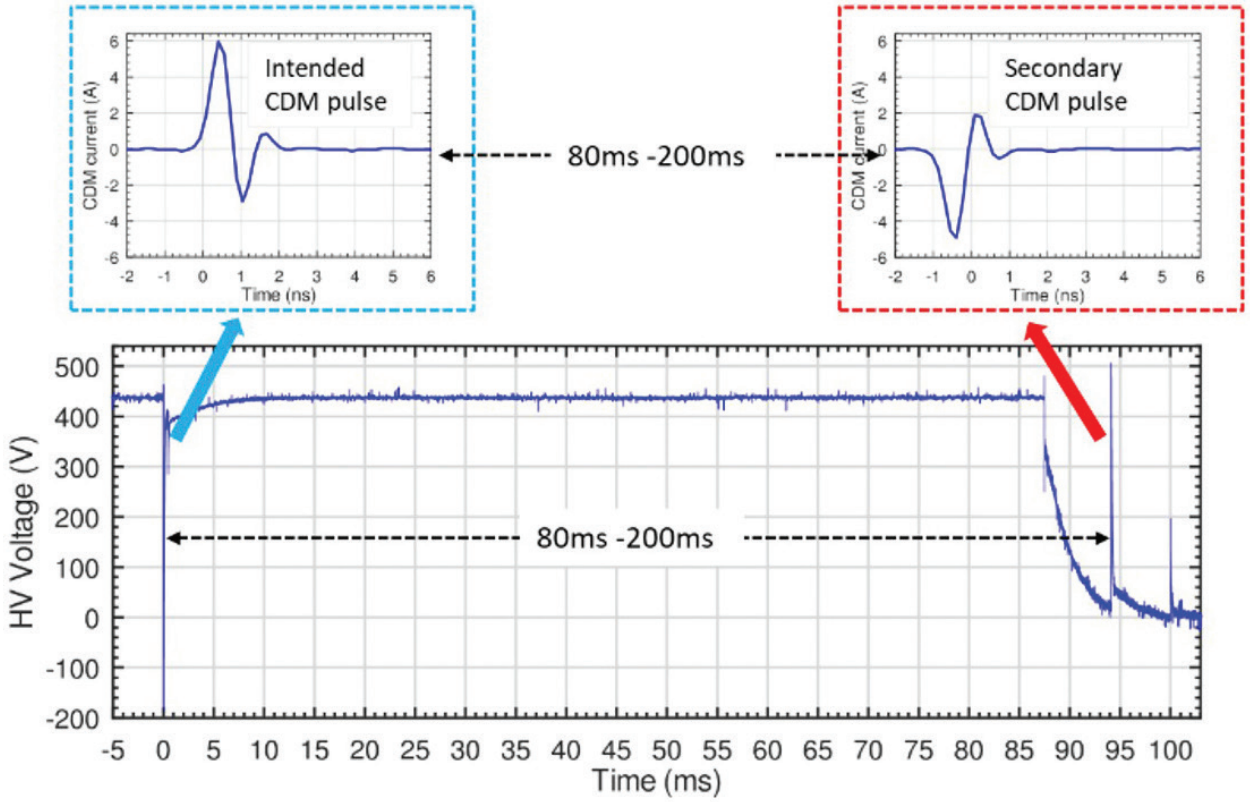

Figure 5: Graph showing field plate voltage and the corresponding CDM current, during a TC +500V

Measurement 2: Electrical measurement of contact bouncing

Measurement set-up to make the breaking events visible by electrical measurements is shown in Figure 6. With this set-up, we expect V_{sense} to read 100mV in case of good contact of pogo pin with DUT and 0V when there is no contact.

Voltage measured at V_{sense} for several Z-stop settings is shown in Figure 7 on page 48. Z-stop is the alignment position in the z-direction. Also marked

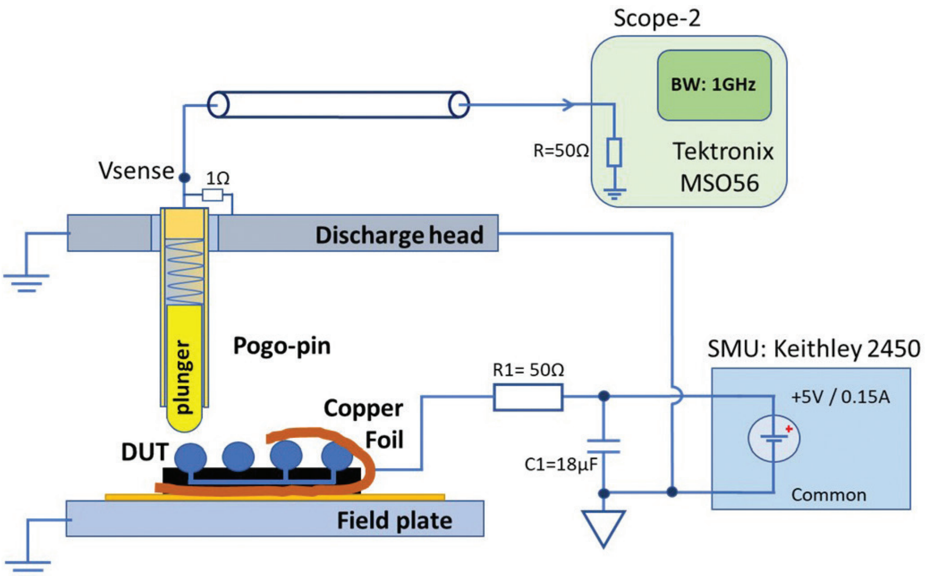

Figure 6: Measurement set-up for monitoring electrical contact between pogo and DUT.

in the graph are the timings of the relevant events. When z = Z-stop, the motor stops movement. The higher the magnitude of z, the deeper the pogo is pushed. From Figure 12-d we see that when Z-stop = 100.96mm, there is no mechanical bouncing. But when Z-stop=100.83mm, the pogo pin does not make contact for most of the time. For any z in between, the contact is broken several times.

CONCLUSION

Secondary discharge is observed when switching of the field plate voltage and mechanical bouncing occurs around the same time. Unfortunately, mechanical vibrations and misaligned pins cannot be totally avoided. To avoid secondary discharge:

1. Increase the overdrive in z-direction. Be careful. too much overdrive can push the sample away.
2. Change tester firmware to ensure switching of the field plate voltage after the vibrations have damped out.

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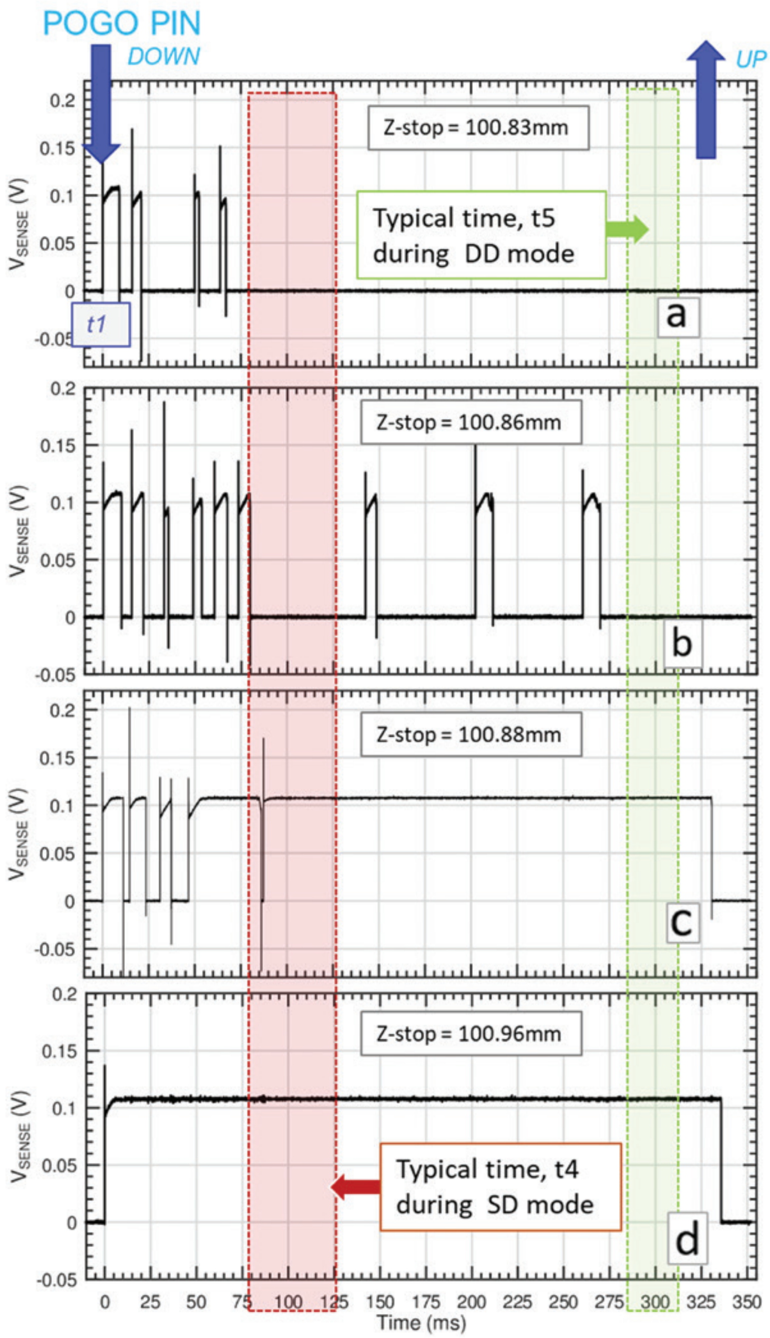
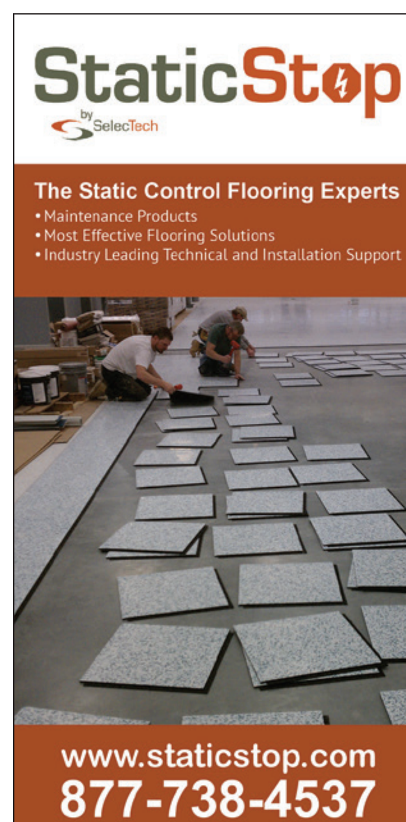


Figure 7: Ohmic contact between pogo-pin and DUT for several Z-stop values

PRODUCT Showcase



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