Fundamentals of DO-160F, Section 22: LIGHTNING INDUCED TRANSIENT SUSCEPTIBILITY

A DASH OF MAXWELL’S
A Maxwell’s Equations Primer - Part 2

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from a European perspective

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I hope you enjoy this January issue. Our authors have done some amazing work this month and we are grateful for their contributions. We also extend thanks to our advertisers who support us with their presence in each issue. And to you, our readers, we thank you for your continued support and feedback. Oh! Don’t forget to vote on whether to include Dilbert in our monthly line-up. We stand ready to follow your wishes, so let us know by entering your vote in our online poll – To Dilbert or Not to Dilbert.

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Fundamentals of RTCA/DO-160F, Section 22
Lightning Induced Transient Susceptibility

by Louis A. Feudi and Robert Given
Thermo Fisher Scientific
Over the past few years, the standard RTCA/DO-160, Section 22 has undergone multiple revisions. As a member of the Aerospace Test Industry, I regularly receive articles and news informing me of the latest changes being implemented to the standard. However, for others who are new to the requirements, many questions are left unanswered. In my travels, I am often asked the same fundamental questions. This article is intended to introduce the requirements of DO-160, Section 22, and to address some of those fundamental questions.

**WHAT IS THE RTCA?**

To better understand the RTCA as it applies to Section 22, let’s take a look at the Foreword for DO-160 Version F.

**WHO IS RTCA?**

The Radio Technical Commission for Aeronautics, organized in 1935 and now known as RTCA, Inc. includes roughly 335 government, industry and academic organizations from the United States and around the world. For a clear understanding of the organization we refer to the Foreword: RTCA, Incorporated is a not-for-profit corporation formed to advance the art and science of aviation and aviation electronic systems for the benefit of the public. The organization functions as a Federal Advisory Committee and develops consensus based recommendations on contemporary aviation issues.[1]

*Author’s Commentary:* This means that the requirements in RTCA DO-160 F are Advisory requirements, not Mandatory requirements.

**WHAT ARE RTCA’S OBJECTIVES?**

Again, let us refer to the Foreword for DO 160F to answer this question. RTCA’s objectives include but are not limited to:

- coalescing aviation system user and provider technical requirements in a manner that helps government and industry meet their mutual objectives and responsibilities;

  *Author’s Commentary:* They mediate requirements between Aircraft part manufacturers, aircraft manufacturers, and Airlines.

- analyzing and recommending solutions to the system technical issues that aviation faces as it continues to pursue increased safety, system capacity and efficiency;

  *Author’s Commentary:* They come up with requirements that try to keep the planes in the air under adverse environmental conditions.

- developing consensus on the application of pertinent technology to fulfill user and provider requirements, including development of minimum operational performance standards for electronic systems and equipment that support aviation;

  *Author’s Commentary:* They get both airplane manufacturers and component manufacturers to agree on a minimum operational performance standard while the product is being stressed (immunity).

  and;

- assisting in developing the appropriate technical material upon which positions for the International Civil Aviation Organization and the International Telecommunications Union and other appropriate international organizations can be based.[1]

  *Author’s Commentary:* Worldwide Organizations adopt these requirements as their positions on issues that arise.

**HOW IMPORTANT ARE RTCA’S STANDARDS?**

Looking again at the Foreword for DO-160F, we learn that the organization’s recommendations are often used as the basis for government and private sector decisions as well as the foundation for many Federal Aviation Administration technical Standard Orders. Since RTCA is not an official agency of the United States Government, its recommendations may not be regarded as statements of official government policy unless so enunciated by the U.S. government organization or agency having statutory jurisdiction over any matters to which the recommendations relate.[1]
The new Thermo Scientific ECAT® Lightning Test System – because failure is not an option.

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**Author’s Commentary:** When an aircraft manufacturer (Boeing, Airbus, DeHavilland, Embraer, Fairchild, General Dynamics, Goodyear, Grumman, Gulfstream American, etc.) is deciding on purchasing criteria for its Tier one, two and three vendor parts, it can use any or all of the requirements in the standard as part of its buying criteria. That means almost all of the electronics incorporated into an airplane need to meet some part of this standard.

**WHO ELSE USES THESE STANDARDS?**

We find the answer to this question also in the Foreword of the standard: These standards were coordinated by RTCA SC-135 with the European Organisation for Civil Aviation Equipment (EUROCAE) Working Groups (WGs) 14 and 31. EUROCAE concurs with RTCA on the environmental conditions and test procedures set forth herein. When approved by EUROCAE, this document will be identified jointly as RTCA DO-160E/EUROCAE ED-14E.[1]

**Author’s Commentary:** The Europeans have similar if not identical requirements, when the standard is adopted as RTCA DO-160E/EUROCAE ED-14E.

**WHAT IS THE SAE?**

SAE International is a global technology information and standards-setting resource for the aerospace, automotive, and commercial vehicle industries. In addition to standards development and publication, SAE holds annual and biennial conferences and tradeshows, periodic industry seminars, student Collegiate Design competitions, all focused on all facets of transportation.

SAE Aerospace is a sub group of SAE International. Their focus is on writing Aerospace Standards (AS), which apply to missile, airframe, ground support equipment, propulsion, propeller and accessory equipment; Aerospace Recommended Practices (ARP), which provide recommendations for engineering design and provide background information and research to support those recommendations: and Aerospace information Reports (AIR), which contain generally accepted industry engineering data and information.

AE-2 is the SAE Lightning Committee established by the Aerospace Council of SAE. They draft and publish the recommended requirements for Indirect Lightning Strikes, based largely on industry and SAE research and knowledge. These requirements are very often adopted into RTCA by the RTCA SC-135 subcommittee, who is responsible for the content of DO-160.

The fundamental document that defined the environment and test waveforms used in DO-160, Section 22, and accounted for the lightning data and analysis necessary to support these requirements, is SAE ARP5412, “Aircraft Lightning Environment and Related Test Waveforms,” originally published in 1999, and revised to Revision A in 2005. This document explains where the concepts of Multistroke, Multi Burst and the current test waveforms originated.

**CHANGES IN REQUIREMENTS OVER THE PAST FEW YEARS**

Many airplane components have to keep working under environmental stress conditions. One of those conditions is lightning. So many airplane manufacturers specify Section 22 as one of the requirements for critical systems, like guidance, radar, communications, engine control, heat and air controls, etc.

---

Figure 1: Current and intended uses for composite materials in the construction of airplanes

Figure 2: A single carbon filament (bottom left to top right) laid across a human hair
We sometimes think of avionics as being the navigation and communications equipment, but it also includes engine controls, servo motor controllers for control surfaces like ailerons, rudder and flaps, landing gear controls, radar, even satellite TV, Wi-Fi, and entertainment systems.

Revisions E and F of DO 160 are driven by the use of composite materials used for airframe construction on recent planes like the Boeing 787 Dreamliner and Airbus A380 (Figure 1).

Carbon Fiber (CF), also known as graphite fiber, or carbon graphite, is a material composed of ultra-thin fibers (0.005-0.010mm diameter) of aligned crystalline Carbon atoms (see Figure 2). The crystalline alignment makes the fiber extremely strong for its size. Thousands of fibers are twisted together to form a yarn, which is often woven into a fabric material. The fiber, combined with a polymer, and heated through various processes, forms a composite material that is extremely strong for its lightweight construction. This lightweight property makes the use of Carbon Fiber Composite (CFC) very attractive to the aerospace industry where weight/thrust ratios are critical for operation, maneuverability, and fuel efficiency.

Prior to CFC materials, the airframe and most other parts of the airplane were made of metal. Thus, if a lightning strike occurred at the nose of the plane, during takeoff for instance, the lightning would travel outside the plane to the tail, exit the surface of the plane, and continue to ground. The solid metal construction of the airframe acted as a Faraday cage, with an extremely low impedance path. This prevented coupling of voltages and currents on the internal wiring of the plane, which usually was routed along the side of the plane, between the inside of the outer skin and the interior bulkhead. This also greatly reduced the susceptibility of the mission critical components located in the plane.

CFC materials, however, don’t conduct lightning currents the way metal airframes do. As a result, the increased impedance of the outer skin as a path for the lightning increases the possibility of higher voltages and currents coupling directly onto internal cables and into the avionics equipment on the aircraft.

**RTCA’s MULTI STROKE, MULTI BURST AND SAE’S ARP5412 REVISION A**

For the newly initiated, DO-160F currently calls out six individual waveforms. Understanding why these waveforms were incorporated into DO-160 comes from referencing the previously mentioned Aerospace Recommended Practice (ARP) 5412 rev. A. Under the scope of the document, the “standardized external current waveforms have in turn been used to derive standardized transient voltage and current waveforms which can be expected to appear on the cable bundles and at equipment interfaces. The test waveforms are considered to be adequate for the demonstration of compliance for the protection of an aircraft and its systems against the lightning environment.” [2]

ARP5412 rev A (hereafter referred to as ARP5412) devotes an entire section to the description of lightning and the variety of forms that lightning can occur. Lightning flashes are the discharging of strong electric fields, or charge centers, within cumulonimbus clouds (Cumulonimbus clouds can form alone, in clusters, or along a cold front in a squall line. They create lightning through the heart of the cloud). There are three types of lightning flashes which may occur:
FEATURE  Lightning Induced Transient Susceptibility

- Flashes between cloud regions of opposite polarity within the same cloud called intra cloud discharges
- Flashes between cloud regions of opposite polarity in different clouds called inter cloud discharges
- Flashes from clouds to ground or the reverse (in instances of high ground locations like mountains and towers)

When a negative cloud to ground flash occurs, the discharge process starts with the formation of a 1-10m wide ionized column that travels in zigzag steps toward the earth. The leader may form branches on the way down to the ground. When the leader gets close to the ground, it causes buildup of high fields near trees and buildings. These then send up leaders to meet the tip of the downward leader. When they meet, a return stroke is initiated, retracing and discharging the leader channel, resulting in a bright flash and high current pulse. After the initial return stroke, subsequent strokes can occur from higher regions of the cloud through the downward leader stroke. These subsequent strokes are usually of lower amplitude than the initial return stroke.

This is the idealized basis for the Multi Stroke Test parameters of DO-160, Section 22. An illustration from the standard is shown in Figure 3. The initial stroke is the highest amplitude of the Multi Stroke application, with subsequent strokes (up to 14) at a lower but repeated amplitude.

Inter- and Intra- cloud discharges behave differently than cloud to ground (and vice-versa) discharges. Most of the data recorded for inter- and intra- cloud discharges come from instrumented aircraft employed in the USA and France to record the characteristics of these types of flashes. Although cloud flashes are less severe than flashes to the ground, it was noted that the rise time of the cloud flashes were significantly shorter (less than 0.4 us) and often occurred in grouped pulses during the initial attachment and final detachment phases of the discharges. These short duration, lower amplitude initial and final phases are the basis for the Multi Burst application of Waveform 3. An illustration of the Multiburst Waveform set is shown in Figure 4. One burst is normally composed of 20 pulses.

WAVEFORMS 1, 2, 3, 4, 5A AND 5B

We now jump to section 7 of ARP5412, which describes the idealized transient waveforms intended for verification of adequate protection of systems and equipment from indirect lightning effects. This section states that there are multiple mechanisms that can induce lightning transients inside the plane from the external lightning environment, but broadly divides them into 2 categories: Aperture Coupling and Resistive Coupling.

Actual induced transients are complex waveforms that result from both coupling methods, but for test purposes, they have been kept separate. ARP5412 states that magnetic fields
Turnkey DO 160F s.22
Lightning Test Solution

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penetrating through apertures (electromagnetically transparent openings) will induce:

1. a current waveshape in conductors or shields terminated to the structure through low impedances at each end (Figure 5)
2. a voltage waveshape in loops existing between cables and the structure (Figure 6)

3. damped oscillatory voltage and current waveforms resulting from excited resonances on coupled interior cables. 1MHz and 10 MHz frequencies representing long and short cable lengths, respectively. (Figure 7)

Resistive Coupling will produce:

1. voltages in conductors within shields due to shield current and shield transfer impedance (Figure 8) and
2. voltages produced by loops existing between cables and the airframe structure (resulting from variances in voltage from the endpoint locations of the cable at different locations on the airframe) and voltages resulting from diffused fields through the structural material (Figure 9). Figure 9 shows 2 different waveforms to cover the wide variance of frame resistance represented by metal and carbon fiber composite airframes.

WAVEFORM POWER LEVELS

There are five Test Power Levels, where Level 1 is the lowest and Level 5 is the highest.

- Level 1
  - Equipment and wiring are installed in a well protected environment
- Level 2
  - Equipment and wiring are in a partially protected environment
- Level 3
  - Equipment and wiring are in a moderate electromagnetic environment
- Level 4 and 5
  - Equipment and wiring are in severe electromagnetic environments

Power levels specified in DO-160 Section 22 are often based upon how critical the component is for flight operation and/or where the unit is located within the aircraft. Greater resistance or spacing from apertures will result in lower power levels. Power levels can also be dictated by the component purchaser or plane manufacturer, based upon other factors. An example is a plane manufacturer may wish to increase the immunity level of entertainment systems on long range aircraft, with the idea that they are more likely to experience multiple lightning strikes and more likely to fail during a 14 hour flight.

TESTING METHODS

There are three primary methods of testing aircraft components

**Pin Injection** – This method is used to directly inject the waveform into connector pins of both cables and printed wiring boards.
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**Cable Bundle Induction** – This method uses a coupling transformer to inductively couple the waveform onto the cable bundle.

**Ground Injection** – This method is often used as an alternate method to inject the waveform onto the ground wire of the unit under test, referenced to the Ground plane that is located on the surface of the test table.

Methods of testing are often dictated by the purchaser, using categorizations defined in DO-160, and involving specific waveform sets for different coupling means. Figure 10 shows the Table for Pin Injection for both aperture coupling and resistive and aperture coupling, and specifies which waveforms are to be applied.

Figure 11 shows the table for Cable Bundle Induction Testing, and indicates the categories for various Aperture and Resistive Couplings. Figure 12 shows an example of a Cable Bundle Injection Transformer.

<table>
<thead>
<tr>
<th>Waveform Set</th>
<th>Test Type</th>
<th>Test Levels</th>
<th>Test Waveform Nos.</th>
</tr>
</thead>
<tbody>
<tr>
<td>C (unshielded, aperture coupling)</td>
<td>Single Stroke</td>
<td>Table 22-3</td>
<td>2, 3</td>
</tr>
<tr>
<td>D (unshielded, aperture and resistance coupling)</td>
<td>Single Stroke</td>
<td>Table 22-3</td>
<td>2, 3, 4</td>
</tr>
<tr>
<td>E (shielded aperture coupling)</td>
<td>Single Stroke</td>
<td>Table 22-3</td>
<td>1, 3</td>
</tr>
<tr>
<td>F (shielded, aperture and resistance coupling)</td>
<td>Single Stroke</td>
<td>Table 22-3</td>
<td>3, 5A</td>
</tr>
<tr>
<td>G (unshielded, aperture coupling)</td>
<td>Single Stroke</td>
<td>Table 22-3</td>
<td>2, 3</td>
</tr>
<tr>
<td></td>
<td>Multiple Stroke</td>
<td>Table 22-4</td>
<td>2, 3</td>
</tr>
<tr>
<td></td>
<td>Multiple Burst</td>
<td>Table 22-5</td>
<td>3</td>
</tr>
<tr>
<td>H (unshielded, aperture and resistance coupling)</td>
<td>Single Stroke</td>
<td>Table 22-3</td>
<td>2, 3, 4</td>
</tr>
<tr>
<td></td>
<td>Multiple Stroke</td>
<td>Table 22-4</td>
<td>2, 3, 4</td>
</tr>
<tr>
<td></td>
<td>Multiple Burst</td>
<td>Table 22-5</td>
<td>3</td>
</tr>
<tr>
<td>J (shielded, aperture coupling)</td>
<td>Single Stroke</td>
<td>Table 22-3</td>
<td>1, 3</td>
</tr>
<tr>
<td></td>
<td>Multiple Stroke</td>
<td>Table 22-4</td>
<td>1, 3</td>
</tr>
<tr>
<td></td>
<td>Multiple Burst</td>
<td>Table 22-5</td>
<td>3</td>
</tr>
<tr>
<td>K (shielded, aperture and resistance coupling)</td>
<td>Single Stroke</td>
<td>Table 22-3</td>
<td>3, 5A</td>
</tr>
<tr>
<td></td>
<td>Multiple Stroke</td>
<td>Table 22-4</td>
<td>3, 5A</td>
</tr>
<tr>
<td></td>
<td>Multiple Burst</td>
<td>Table 22-5</td>
<td>3</td>
</tr>
</tbody>
</table>

As defined in Figures 10 and 11:

a. A, C, E, G and J are for aperture coupling
b. B, D, F, H, and K are for resistance coupling
c. A and B specify Pin Injection
d. C through F specify cable bundle single stroke
e. G through K specify cable bundle single, multi-stroke and multi-burst
f. Z means other tests were conducted

These letters (A-K, Z) are used as part of a classification code often provided by the buyer (Airplane Manufacturer) who determines what coupling may occur and what induced conditions may be expected on their airplane, based upon construction materials, unit location and expected coupling zones.

**MIL STD 461**

The content presented above represents an explanation of the fundamentals of DO-160 Section 22 Testing for Indirect Lightning effects. Understanding the origins and justifications of the requirements often help in determining the course of action needed to test a product. Current expectations are that the requirements listed in DO-160 Section 22 will be adopted by MIL STD 461 in 2011.

**REFERENCES**

1. RTCA/DO-160F Environmental Conditions and Test Procedures for Airborne Equipment
2. SAE ARP5412-RevA-2005 Aircraft Lightning Environment and Related
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Robert Given has over 6 years experience in the Compliance Testing Field working for Thermo Fisher Scientific as Project Manager, Engineering Manager and now Application/ Product/ Technology Manager. Prior to joining Thermo Fisher, Bob was Principal Engineer at Kokusai Semiconductor Equipment Corporation designing and managing control systems for Diffusion and Chemical Vapor Deposition semiconductor equipment for 25 years. Bob can be reached at bob.given@thermo.com or (978) 935-9333.

**Figure 12: Injection Transformer**
For Waveforms 1, 5A and 5B

*Photo courtesy of Thermo Fisher Scientific*
A Dash of Maxwell’s Equations Primer

Part 2
Why Things Radiate

In Chapter I, I introduced Maxwell’s Equations for the static case, that is, where charges in free space are fixed, and only direct current flows in conductors. In this chapter, I’ll make the modifications to Maxwell’s Equations necessary to encompass the “dynamic” case, that is where magnetic and electric fields are changing. Then I will try to explain why things radiate.

Here are Maxwell’s equations for the static case:

\[
\begin{align*}
\oint \mathbf{D} \cdot d\mathbf{s} &= \iiint \mathbf{E} \cdot d\mathbf{s} = Q \\
\oint \mathbf{B} \cdot d\mathbf{s} &= \iiint \mathbf{H} \cdot d\mathbf{s} = 0 \\
\frac{1}{\varepsilon_0} \oint \mathbf{D} \cdot d\mathbf{l} &= \oint \mathbf{E} \cdot d\mathbf{l} = 0 \\
\frac{1}{\mu_0} \oint \mathbf{B} \cdot d\mathbf{l} &= \oint \mathbf{H} \cdot d\mathbf{l} = I
\end{align*}
\]

Where:

- \( \mathbf{D} \) = Electric flux density = \( \varepsilon_0 \mathbf{E} \)
- \( \mathbf{E} \) = Electric field in volts/meter
- \( \mathbf{B} \) = Magnetic flux density = \( \mu_0 \mathbf{H} \)

Figure 1: Kirchhoff’s voltage law is illustrated in (a). The voltage around a closed loop is zero. In (b), a changing magnetic flux introduces an additional time varying voltage across the resistor.
$H = \text{Magnetic field in amps/meter}$

$\varepsilon_0 = \text{Free space permittivity} = 8.85 \times 10^{-12}$

$\mu_0 = \text{Free space permeability} = 4\pi \times 10^{-7}$

The first of the modifications we need to explain the “dynamic” case we owe to the work of Michael Faraday. For the static case, the third equation states that the electric field integrated around a closed loop (the “line integral”) is zero.

$$\oint E \cdot dl = 0$$

Engineers more commonly deal with this in the form of Kirchhoff’s voltage law:

$$\sum_{\text{Closed-Circuit}} V = 0$$

Faraday’s contribution was to establish that Kirchhoff’s voltage law is nearly always wrong. Where there is a changing magnetic flux through a loop, a voltage is created by that changing flux. That voltage is equal to:

$$\oint E \cdot dl = V = \frac{\partial B}{\partial t} \cdot A$$

Where $A$ equals the area of the loop.

The effect of this flux-induced voltage is illustrated in Figure 1(b). It shows up as an additional time varying voltage across the resistive load. So to account for changing magnetic fields through the loop, we must modify Maxwell’s third equation as follows:

$$\oint E \cdot dl = \frac{\partial B}{\partial t} \cdot A$$

This equation explains those ever-present and annoying “ground loops.” They can be minimized by minimizing either the strength of the magnetic field ($B=\mu_0 H$), its rate of change ($\partial B/\partial t$) or the area of the loop ($A$). The equation assumes that the loop is two dimensional and the field uniform across the loop at any given instant. Where neither is so, we need a more generalized solution:

$$\oint E \cdot dl = \int \int \frac{\partial B}{\partial t} \cdot ds$$

This equation is known as the “integral form” of Maxwell’s third equation, but it’s cumbersome to use, and, for the most part, we’ll be dealing with two dimensional loops and fields that at any given instant are uniform over the loop area, so we can work with the simpler form.

It was Maxwell himself who completed what was to become the fourth of his equations for the dynamic case. The fourth equation for the static case states:

$$\oint \frac{1}{\mu_0} B \cdot dl = \oint H \cdot dl = I$$

The problem lay with the definition of current, $I$. Today, engineers are comfortable with thinking of current traveling through circuits either by way of conduction, by capacitive coupling or by induction. Faraday dealt with induction. Maxwell’s contribution was to separate “conduction” current from “capacitive” current, the latter which he called “displacement” current.

$$\oint H \cdot dl = I_{\text{conduction}} + I_{\text{displacement}}$$

Let’s take the case of a parallel plate capacitor where $C=\frac{A\varepsilon_0}{d}$, $E=\frac{V}{d}$, and $Q=CV$. Noting that by definition, the time derivative of charge equals the current ($I=\frac{dQ}{dt}$) the displacement current passing through a parallel plate capacitor is equal to:

$$Q = CV$$

$$\frac{dQ}{dt} = C \frac{dV}{dt}$$

$$\frac{dQ}{dt} = I_{\text{displacement}}$$

$$C = \frac{A\varepsilon_0}{d}$$

$$V = E \cdot d$$

$$I_{\text{displacement}} = A\varepsilon_0 \frac{\partial E}{\partial t}$$

Combing the above yields:

$$\frac{1}{\mu_0} \oint B \cdot dl = \oint H \cdot dl = I_{\text{cond}} + A\varepsilon_0 \frac{\partial E}{\partial t}$$

Variable $A$ is, of course, the area of our capacitor’s plates. As long as we’re dealing with parallel plate capacitors, we can...
Our products have always outlasted and outperformed the competition, now we’re giving them another problem.

We’ve shrunk our “S” amplifiers giving you more power with an even greater price-performance ratio.

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Our competitors have some other choice words for it. But that’s their problem.

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In Europe, call at United Kingdom 441-933-26766 • at France 33-1-47-91-75-30 • e-mail: ar GmbH 89-614-1710 • at Benelux 31-172-423-000

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use the equation in the form above. More generally, however, area can be expressed as:

$$A = \iint f(s) \cdot ds$$

Here, \( f(s) \) is a function that is integrated over a surface (or an envelope) to calculate the flux. In our case the function \( f(s) \) is the time derivative of the electric field density, \( D \), \( f(s) = \partial D/\partial t \), so:

$$\frac{1}{\mu_0} \oint B \cdot dl = I_{cond} + \iiint \frac{\partial D}{\partial t} \cdot ds$$

We can now state all four of Maxwell’s equations in general form:

\[
\begin{align*}
\iiint D \cdot ds &= Q \\
\iiint B \cdot ds &= 0 \\
\frac{1}{\varepsilon_0} \oint E \cdot dl &= -\iiint \frac{\partial B}{\partial t} \cdot ds \\
\frac{1}{\mu_0} \oint H \cdot dl &= I_{cond} + \iiint \frac{\partial D}{\partial t} \cdot ds
\end{align*}
\]

Somewhat more intuitively, we can state Maxwell’s Equations in words:

1. The electric flux through a closed envelope equals the charged contained.
2. The magnetic flux through a closed envelope is zero.
3. The electric field integrated around a closed loop (the “line integral”) equals the negative of the rate of change of the magnetic flux through the loop.
4. The magnetic field integrated around a closed loop is equal to the total current, both conductive and capacitive, that passes through it.

Next, I'll try to explain why radio waves radiate.

First, I'll have to take you to a place far, far away where there are no conduction currents and no free charges, a place we can truly call free space. There, Maxwell’s Equations reduce to the following:

\[
\begin{align*}
\iiint D \cdot ds &= 0 \\
\iiint B \cdot ds &= 0 \\
\oint E \cdot dl &= -\mu_0 \iiint \frac{\partial H}{\partial t} \cdot ds \\
\oint H \cdot dl &= \varepsilon_0 \iiint \frac{\partial E}{\partial t} \cdot ds
\end{align*}
\]

In free space, we need only to deal with the third and fourth of Maxwell’s equations:

\[
\begin{align*}
\oint E \cdot dl &= -\mu_0 \iiint \frac{\partial H}{\partial t} \cdot ds \\
\oint H \cdot dl &= \varepsilon_0 \iiint \frac{\partial E}{\partial t} \cdot ds
\end{align*}
\]
Our next task is to find expressions for the electric and magnetic fields that satisfy these two equations. I’ll do this using a time honored tradition in calculus. I’ll guess at the answer and then plug the answers into the equations to see if they work. Figures 2 and 3 show my guesses.

My proposed solution is a set of two fields, set perpendicular to each other as shown in Figure 4. Figure 2(a) shows the electric field at time equals zero. The electric field vector points in the z direction and it varies sinusoidally with time. As such, the entire waveform appears to move in the direction the x direction. Mathematically, it is expressed as:

\[ E_z = -E_0 \sin(\omega t - kx) \]

Where:
- \( \omega = \) The frequency in radians per second = \( 2\pi f \)
- \( f = \) The frequency in Hertz
- \( k = \) The “wavenumber” = \( 2\pi/\lambda \)
- \( \lambda = \) the wavelength in meters

Likewise, the magnetic field is oriented in the y direction and it also appears to move in the x direction. It is expressed as:

\[ H_y = H_0 \sin(\omega t - kx) \]

Figure 4 shows this combination which is known as a “plane wave.” The crests of the magnetic and electric fields seem to move through space in the positive x direction as if they were a wall, hence the term plane wave.

Next, we’ll plug the proposed solution for the electric field into the third of Maxwell’s Equations. To do this we’ll have to calculate a line integral. The line integral is equal to the field times the distance around a closed loop.

Fortunately, there is an easy, graphical way to calculate the line integral. What we want to do is to find a convenient loop and multiply the field times the perimeter of the loop. The location that we pick for our convenient loop is shown in Figure 5(b). The loop aligns on the left with location \( x_0 \) and on the right with \( x_1 \).

At \( x_0 \), the electric field is at its maximum and is equal to \(-E_0\). At a slight distance to the right, \( x_1 \), the electric field has lessened in magnitude slightly. At \( x_1 \), amplitude is:

\[ E_{x1} = -(E_0 + \left. \frac{\partial E}{\partial x} \right|_{x_1 - x_0}) = -(E_0 + \left. \frac{\partial E}{\partial x} \right|_{\Delta x}) \]

In order to preserve the right hand rule, which requires us to move in a counter clockwise direction, we’ll begin our line integral calculation by a move of a distance \(-\Delta z\) as shown in Figure 5(b), creating the first component of our loop integral. This first component is equal to \((-E_0)(-\Delta z) = E_0 \Delta z\). There’s no electric field in the x direction, so we don’t have to consider the top and bottom sides of our rectangular loop. On the right side of our loop, we move a distance \( \Delta z \) times the field at that point. Adding the contributions of our loop movement together and noting that \( \Delta x \Delta z \) equals the area of the loop (A), we get:

Figure 3: This proposed solution for Maxwell’s Equations uses, as its other component, a magnetic field as shown. It is time correlated with the electric field of Figure 2, but is oriented 90 degrees from it in space.
Substituting this expression for the line integral of the electric field, and noting that:

\[
\oint E \cdot dl = \oint E \cdot d\ell = E_0 \Delta z + \left( - \frac{\partial E}{\partial x} \Delta y \right) + \left( - \frac{\partial E}{\partial y} \Delta x \right) = E_0 \Delta z - \frac{\partial E}{\partial x} \Delta x \Delta z = \frac{\partial E}{\partial x} A
\]

We find that:

\[
\oint E \cdot dl = \oint \frac{\partial E}{\partial x} ds = A
\]

We can do the same for magnetic fields deriving a similar equation:

\[
\frac{\partial H}{\partial x} = \varepsilon_0 \frac{\partial E}{\partial t}
\]

It’s now time to plug in the proposed solution -- the plane wave -- to see if it works. The proposed solutions was:

\[
E_z = -E_0 \sin(\omega t - kx)
\]

\[
H_y = H_0 \sin(\omega t - kx)
\]

Taking the derivative of \(E_z\) and \(H_y\) with respect to time and distance yields:

\[
\frac{\partial H_y}{\partial t} = -k H_0 \cos(\omega t - kx)
\]

\[
\frac{\partial E_z}{\partial t} = -k E_0 \cos(\omega t - kx)
\]

\[
\frac{\partial H_y}{\partial x} = \omega H_0 \cos(\omega t - kx)
\]

\[
\frac{\partial E_z}{\partial x} = \omega E_0 \cos(\omega t - kx)
\]

Therefore:

\[
\frac{\partial H_y}{\partial x} = \frac{dE_z}{dt}
\]

\[
\frac{\partial E_z}{\partial x} = \frac{dH_y}{dt}
\]

\[
-k H_0 \cos(\omega t - kx) = \varepsilon_0 \omega E_0 \cos(\omega t - kx)
\]

\[
-k E_0 \cos(\omega t - kx) = \mu_0 \omega H_0 \cos(\omega t - kx)
\]

\[
H_0 = \frac{\varepsilon_0 \omega}{k} E_0
\]

\[
-k E_0 = \mu_0 \omega H_0 = -\mu_0 \omega \frac{\varepsilon_0 \omega}{k} E_0
\]

\[
\frac{1}{\sqrt{\mu_0 \varepsilon_0}} = \frac{\omega}{k}
\]

The proposed solution works if \(1/\sqrt{\mu_0 \varepsilon_0} = \omega/k\). Does it? Note that \(\omega = 2\pi f\), \(k = 2\pi/\lambda\) and therefore \(\omega/k = f\). The units of \(f\) are cycles/second times meters/cycle, or meters/second = velocity. The term \(\omega/k\) must be the velocity of the wave as it moves in the x direction.
As for $1/\sqrt{\mu_0\varepsilon_0}$, it is equal to:

$$\frac{l}{\sqrt{l/\mu_0\varepsilon_0}} = \frac{l}{\sqrt{4\pi \times 10^7 \times (8.85 \times 10^{-12})}} = 3 \times 10^8 \text{ meters/second}$$

This, of course, is the speed of light ($c$), which is exactly what we would expect.

Before closing this chapter, we’ll use the equations above to derive two characteristics of plane waves. Since the electric field is expressed in terms of V/m and the magnetic field in A/m, dividing E by H at any given point in space produces a resultant that is in units of V/A, or Ohms. In free space, this ratio is:

$$\frac{E_z}{H_y} = \frac{k}{\varepsilon_0 \omega} = \frac{l}{c \varepsilon_0} = \frac{\mu_0}{\varepsilon_0} = 377 \text{ ohms}$$

The “impedance” of free space, we can conclude, is 377 ohms.

We can also multiply the magnitudes of the electric and magnetic fields at any point in space yielding a resultant that is in units of V/meter x A/meter or Watts/meter$^2$. From that we can conclude that a plane wave transmits power in the direction of its motion.

$$P_x = E_z \times H_y$$

P is known as the Poynting vector.

Why do things radiate? In short, electromagnetic fields radiate because a change in the electric field with time causes a change in the magnetic field around it. That, in turn, causes a change in the magnetic field with time which causes a change in the electric field around it. The two fields alter each other, causing a movement through space over time. ■

Glen Dash is the author of numerous papers on electromagnetics. He was educated at MIT and was the founder of several companies dedicated to helping companies achieve regulatory compliance. Currently he operates the Glen Dash Foundation which uses ground penetrating radar to map archaeological sites, principally in Egypt.

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HEADQUARTERS NEWS
We begin 2010 with great optimism. Last year, 2009, our certification renewals and applications for new certifications exceeded our expectations. We are delighted to see that the value of our credential has been so well appreciated during these difficult times for our global communities.

All our new programs and initiatives announced in 2009 started well, and our Education Advisory Committees, EACs, are already setting dates and venues for both Professional Development Workshops and iNCLA Credentialing Workshops to be held in 2010.

I am not sure what we did but our neighbors decided to move out a few months ago. Maybe it was something we said!!!

Or maybe they heard about the musical talents of some of the iNARTE Board.

NEW PROGRAM PLANS FOR 2010
iNARTE Certified Laboratory Auditor, iNCLA
This new program to certify those internal auditors, who have the responsibility to prepare a Test or Certification laboratory for assessment to ISO/IEC 17025, was launched in 2009. Two training and credentialing workshops are now planned for 2010 in cooperation with the ANSI-ASQ National Accreditation Board, ACLASS. The first will be held in Atlanta, March 16th-19th, with the second being in Chicago, September 14th-17th. Watch for detailed venue announcements on the iNARTE website, www.narte.org.

Associate Engineer and Associate Technician. iNAE/iNAT
Associate certification is now available to graduates from iNARTE accredited Universities and Training Institutes who achieve a high GPA and who are endorsed by a senior member of faculty. This credential enables new graduates to enter the work force and build their career, while enjoying the advantages of iNARTE membership as they accumulate experience for full Certification. Graduates from other curricula may also apply for this credential, but further examination will be required.

This new iNARTE program has been enthusiastically received by both the IEEE EMC Society and the IEEE Product Safety Engineering Society. Both have agreed to offer all iNARTE Associates a one year free membership with all the associated benefits this offers to young engineers and technicians starting their careers.

The first University in the USA to become iNARTE accredited was Missouri University of Science and Technology, MUST, who offer a range of EMC and ESD courses, including hands on training at their very well equipped EMC Laboratory.

We did not realize it was a mobile home next door
Training and Professional Development

INARTE’s Education Advisory Committees are busy planning their 2010 schedule of professional development workshops. It is expected that by popular demand both of the 2009 events will have to be repeated in 2010, with the addition of at least two others. The two repeat events will be:

ANSI C63.10 Workshop – Expected to be at UL Northbrook, IL

High Power Electromagnetic (HPEM) Threats – Expected to be in Gaithersburg, MD

Watch the INARTE website, www.narte.org, for specific details on these and our other events as they are developed.

Brian Lawrence, INARTE, presents the MUST Accreditation Certificate to students at the Universities EMC laboratory

Elya Joffe and Mike Violette, both members of the INARTE Board, entertain attendees at EMC2009 in Austin
INTRODUCTION
Do you supply products into Europe? If you supply products that come within the scope of the EMC Directive 2004/108/EC, the application of harmonized standards provides the simplest means of demonstrating conformity with the protection requirements (emission and immunity) of the Directive.

This article will provide you with essential information on the selection and use of the appropriate standards for your product.

HOW DO STANDARDS AND DIRECTIVES INTERACT?
Directives such as the EMC Directive are so called “new approach” directives. These were introduced from 1985 onwards as a means of speeding up the creation of technical requirements that could be applied throughout Europe. Before that date the regulations contained all the technical requirements for products within their scope. Agreeing on these requirements was a lengthy process, and the legislation was inflexible, incapable of responding quickly enough to technical innovation.

New approach directives set out only the essential requirements in general terms; the technical requirements are contained in harmonized standards that underpin them. However, an important feature of new approach directives is that the use of harmonized standards is always voluntary. The manufacturer can demonstrate conformity with the essential requirements by other means provided that they justify their approach in technical documentation that shows the technical analysis that they have followed.

Most manufacturers apply the requirements of the harmonized standards because they define the technical requirements clearly and are the equivalent of carrying out the electromagnetic compatibility assessment required by the EMC Directive. They also have the advantage that compliance in full with the requirements of all the relevant harmonized standards provides a “presumption of conformity” with the essential requirements of the directive. This means that the market surveillance authorities must presume that a product that is stated to meet the requirements of the harmonized standards meets the technical requirements of the directive, and they cannot remove products from the market unless they can demonstrate that a product does not comply. Where the harmonized standards have not been applied in full, this presumption does not exist.
WHAT IS A HARMONIZED STANDARD?

Europe has a series of standards prefixed “EN” - European Norm. These are written by the European Committee for Standardization (CEN), the European Committee for Electrotechnical Standardization (CENELEC), and the European Telecommunications Standards Institute (ETSI). The vast majority of ENs that are relevant for the EMC Directive are produced by CENELEC.

Harmonized standards are ENs produced by CEN, CENELEC or ETSI, following a mandate issued by the European Commission, for use with one or more directives. The lists of harmonized standards suitable for each Directive are published from time to time in an official publication called the Official Journal of the European Union, often referred to as “the Official Journal” or “the OJ”.

This article will describe the application of the standards produced by CENELEC to products in order to meet the requirements of the EMC Directive. The principles for the standards produced by CEN are similar. ETSI standards are produced entirely in Europe and do not have international equivalents.

WHY DOESN’T EUROPE USE INTERNATIONAL STANDARDS PUBLISHED BY IEC?

The majority of ENs are not written exclusively in Europe, but are based on international standards; in many cases the technical content is identical. In the case of CENELEC in 2008, 78.1% of standards published were identical to IEC standards (including CISPR) and a further 4.5% were based on IEC standards (including CISPR), with some modification for European requirements.

An agreement between CENELEC and IEC ensures that any new standardization work required to be produced by CENELEC is first of all offered to IEC, to allow an international version to be produced. CENELEC then takes the text of the international standard in order to produce the EN. Only if IEC rejects the offer, or cannot complete the work to required deadlines, does CENELEC work on the production of the standard.

European legislation is constructed in a manner that requires harmonized standards setting out the technical requirements of a directive to be voted on as European standards. As a consequence, international standards are subject to two votes in Europe, once as the international standard and once as the EN. If both votes are positive, two versions of the same standard are published, the IEC and the EN.

If the international vote is positive, but the voting in Europe shows that there is insufficient support for publication as an EN, then changes are made to the requirements in the EN until a positive vote is achieved. These differences are called “common modifications”. Common modifications may consist of addition of, deletion of, or changes to requirements or test methods.

Another reason why the EN version is published is that IEC standards are recommendations that come into effect immediately; the superseded standard is withdrawn when the new version is published. If this were to have the legal status of a harmonized standard, the change in requirements would be too abrupt to allow industry time to adapt their products to the new requirements. The European versions have a (usually) three year transition period from the date the standard is ratified to the date that the superseded standard is withdrawn. This period applies to all types of standards, and the date of withdrawal of the superseded standard is published inside the front cover of the EN standard. In the case of harmonized standards listed in the Official Journal, the date published in each list in the OJ becomes the relevant date in respect of the application of the standard under that particular directive (see below).

TYPES OF STANDARDS

EMC standards are of several different types: product, product family, generic and basic. Product and product family standards define the requirements and test methods for a small range of products. Generic standards define the requirements and test methods for those product types that are not covered by the more specific product and product family standards. Generic standards are based on types of environment rather than product categories. Finally, basic standards set out test methods or provide guidance and background information. They may contain recommendations but do not set absolute requirements. Consequently, basic standards do not of themselves provide a presumption of conformity. Rather they provide standardized test methods that can be referenced from the other standard types.

HOW TO SELECT HARMONIZED STANDARDS - THE LIST IN THE OFFICIAL JOURNAL OF THE EUROPEAN UNION

The latest list of harmonized standards that provide a presumption of conformity under the EMC Directive (at the time of writing this article) is available in PDF format at this link (www.incompliancemag.com/link/1001_100) although there has been a change in the date in respect of the 1998 edition of EN 55022 (www.incompliancemag.com/link/1001_101).

It will be seen that there are four columns in the table, and that there is an entry for each standard, with any amendments being identified separately. The first column identifies the standardization body that publishes the standard. The second column provides the number and title of the standard.
the standard is based on an international standard the number is shown in brackets below the title. Where the EN contains common modifications, “modified” is shown.

The third and fourth columns deal with the editions of the standards and amendments that provide a presumption of conformity. For each standard or amendment, the third column lists the standard that is superseded by that standard or amendment. Often, in the case of a new edition of the standard, the superseded standard is the previous edition of the standard and any amendments to it, but it can be the relevant generic standards where a new standard is published that covers products that were not previously within the scope of a product or product family standard.

Amendments are dealt with separately, so for example the superseded standard in the case of an amendment 2 is the standard with its amendment 1. Amendment 2 of the standard has to be applied with effect from the date given for the presumption of conformity to be valid.

The fourth column, entitled “date of cessation of presumption of conformity of the superseded standard” provides the relevant dates. The date is chosen by the Commission; it is generally the same as the date associated with the three year transition period set by CENELEC, but is some exceptional cases it is different. The date listed in this column is definitive in respect of the legal position on whether the correct edition of a standard has been selected. In cases where the date of cessation is earlier than the date that the list was published, it is shown as “date expired” and the relevant date is shown in brackets.

Notes to the list explain the requirements in particular circumstances and in contrast to notes in standards, these notes have legislative effect.

**DETERMINING WHICH STANDARDS ARE RELEVANT**

The selected harmonized standards should make a complete provision for emissions and immunity, at both high and low frequencies. In selecting standards from the list in the OJ, the manufacturer should be aware that more specific standards take precedence over more general standards and that the harmonics and flicker standards, EN 61000-3-2 and EN 61000-3-3 respectively, apply to all products intended for connection to the public low-voltage mains electricity supply.

In general, the scopes of the product standards are mutually exclusive. However convergence of functionality is creating products for which requirements are not complete within existing standards. New standards such as CISPR 32 and CISPR 35 are being developed for emissions and immunity, respectively, of multimedia equipment, and these will be harmonized in Europe as EN 55032 and EN 55035. Until such time as more comprehensive standards are available, it may be necessary to apply the parts of more than one standard for each aspect. For example, a computer with a broadcast television reception function will come within the scope of EN 55022 and EN 55024 as an information technology product, but parts of EN 55013 and EN 55020 will need to be applied in respect of the broadcast reception functions, in order to make a complete assessment of the EMC performance for all aspects of the equipment. This still represents a simpler approach for the manufacturer than carrying out an electromagnetic compatibility assessment of the product in accordance with Annex II, point 2 of the EMC Directive.

Although the titles of harmonized standards can be a useful indication of appropriate standards, it is often necessary to examine the scope and even the content of a standard to check its applicability to a particular product. For example, the harmonized standard EN 55014-1 is entitled *Electromagnetic compatibility — Requirements for household appliances, electric tools and similar apparatus* yet within its scope are the following (non exhaustive list): Cine projectors, automatic dispensing (vending) machines, juke boxes, pinball machines, gaming machines with a winnings-payout mechanism, electric fences, cow milking machines, and air conditioners.

The complete list should be examined in its entirety for more specific standards that would take precedence. For example, EN 55022 and EN 55024 apply to information technology.
FEATURE  EMC Standards from a European Perspective

WHICH EDITIONS OF THE STANDARDS CAN/MUST BE USED?

In the case of EMC harmonized standards listed in the Official Journal, the standard provides a presumption of conformity with the relevant essential requirements as soon as it is included in the list, and published by at least one national standards body. This may be some months after the standard is actually published by national standards bodies. During the period up to the date of cessation of presumption of conformity of the superseded standard, both the new and superseded editions provide a presumption of conformity. The superseded standard ceases to provide this presumption on the date given, so manufacturers using the standard should update their declarations of conformity accordingly, having first satisfied themselves that their product complies with the requirements of the new edition.

Many harmonized standards refer to basic standards for test methods, and these references may be dated or undated. In the case of dated references the specific edition of the referenced standard, including any amendments specified, must be used, even if a later edition of that standard has been published. With undated references, the latest edition of the referenced standard must be used, although the previous edition may be used up to its date of withdrawal. A dated reference provides more certainty for the manufacturer, and the majority of new EMC standards being produced by the main EMC committees are now adopting dated references. This may not be the case, however, for standards that contain EMC requirements along with other requirements that are produced by the product standards committees.

A special case occurs where the referenced standard is itself a harmonized standard listed in the Official Journal. This can occur where for historical reasons a product family standard defines test methods within the body of the standard because suitable basic standards were not available at the time that the first edition of the product family standard was written. Another harmonized standard then makes reference to the product family standard.

The principle to be followed in this case, where the reference is dated, is that the standard for the product in question is the primary requirement, and that the edition of the referenced standard that is to be used is the one determined by the product standard. Thus, the dated reference is followed, irrespective of whether that edition is listed in the currently valid list in the Official Journal.

WHAT IS ANNEX ZA IN THE EUROPEAN STANDARDS?

When an international standard is ratified by CENELEC as an EN, the national standards bodies that publish the standard take the entire international text (subject to any common modifications that have been necessary). This means that throughout the text, references to other standards will be to their international versions. For Europe, it is the EN versions of these standards that are the relevant references, and Annex ZA has been developed to deal with this. It is so designated to ensure that it comes after the normative and informative annexes of the international text.

Annex ZA consists of five columns. The first two provide the number and date of the international version of the standard as referenced in the international body text of the standard. The third column provides the title of the referenced document. The third and fourth columns provide the number and date of the corresponding European version of the referenced standard that must be used for the correct application of the standard. Where no equivalent is published a dash appears in the right hand columns, and the internal version must be used.

If a year is not given in the fifth column, the reference is undated. It is possible for the international version to be an undated reference and for the EN reference to be dated.

It is recognized that the wording of Annexes ZA as currently provided is not entirely clear, and work is in hand in Europe to improve this situation.

ANNEX ZZ

The informative Annex ZZ in harmonized standards is a relatively recent innovation that is being introduced as standards are amended or new editions are published. It seeks to provide guidance to users of the standards on the coverage in the standard of essential requirements under various directives. It should be used with care, because it does not indicate whether the standard makes a complete provision in respect of the essential requirements that it identifies, or whether other standards must be applied in addition. For example, EN 55022 contains an Annex ZZ that indicates that it covers EMC emission aspects for the EMC Directive 2004/108/EC and the R&TTE Directive 1999/5/EC. However it does not indicate that the harmonics and flicker standards are applicable, in addition, for equipment within its scope that is connected to the public low-voltage electricity supply.
**INTERPRETATION SHEETS**

In some cases, a requirement in a standard is found to be unclear or ambiguous, yet the problem is not sufficient to warrant an immediate amendment to the standard. Such cases are dealt with by an Interpretation Sheet. These are a relatively new development, and may be published to clarify a requirement in a standard, but not to modify it; such changes would be the subject of an amendment.

In the British Standard version of ENs, Interpretation Sheets are included as amendments to the standard, but this is not the case for all national implementations of ENs. The existence of interpretation sheets may be checked in the online catalogue on the CENELEC website www.cenelec.eu, where putting the number of the standard (without “EN”) in the search box on the top right hand side of the home page will produce a list of current and draft documents associated with that standard.

In respect of the application of a standard and the presumption of conformity under a directive, Interpretation Sheets are guidance, and therefore such documents referring to harmonized standards are not listed in the Official Journal. The presumption of conformity is not affected if they are not followed, but since they represent the official opinion of the responsible standards committees, it would be prudent to follow the interpretation.

**ALTERNATIVE TEST METHODS**

Many harmonized EMC standards offer alternative test methods for demonstration with the requirements. The question has arisen in recent years as to how these should be considered, especially if a challenge is made by market surveillance authorities to a declaration of conformity under the EMC Directive that is supported by compliance with such standards.

The subject has proved to be controversial, but the interpretation in Europe is clear, resulting in a statement in the European Commission’s *Guide for the EMC Directive 200/108/EC*. This states “alternative test and measurement methods, when introduced into a harmonized standard for the same purpose are considered, together with their associated limits, as equivalent regarding the provision of a presumption of conformity with the protection requirements”.

European standards are being checked for consistency with this position, and in some cases this results in a change (by common modification) from the international equivalent. The wording in EN 55022:2006 dealing with this issue reads as follows: “Where alternative test methods are described in the following subclauses, compliance with the requirements of the subclause may be demonstrated by either or any of the methods described.”

**DOCUMENTS PROVIDING FURTHER GUIDANCE**

CENELEC has produced Guides 24 and 25 to provide explanations of the use of harmonized standards for EMC. They may be downloaded free of charge from the following web page http://www.cenelec.eu/Cenelec/CENELEC+in+action/default.htm.

Guide 25 *Guide on the use of standards for the implementation of the EMC Directive* will be the most useful for manufacturers, although Guide 24 *EMC standardization for product committees* also provides useful background information.

The third editions of these guides are about to be published, and may be available by the time this article appears in print. These update the references to the new EMC Directive 2004/108/EC and to other documents.

Brian Jones is an independent EMC Consultant, specializing in compliance with European legislation and standards. He is also secretary to the CENELEC EMC committee TC210, but is writing here in a personal capacity, and his views do not necessarily reflect the views of any organization. He may be contacted at emc@brianjones.co.uk.
Time-Saving Effects of FFT-Based EMI Measurements

by Vic Hudson
Rohde & Schwarz
In the world of RF and microwave testing, measurements required for EMI are among the most complex and time-consuming since they incorporate a wide array of specific tests that must be performed over an array of frequencies. They typically require not only many hours of test time but even more for configuring and reconfiguring the test set-up.

Fortunately, advances in the signal processing abilities of test equipment have reduced test time over the years. However, the real improvements are the result of enhancement measurement software, greater integration, automation of the test process, and increasing acceptance of time-domain techniques based on Fast Fourier Transform (FFT) for use in preview measurements of the disturbance spectrum, for example. Together they are slowly making the EMC measurement process faster, and more efficient and accurate.

Over the years, there has been a continuing trend toward greater automation of test environments that includes more fully integrating the elements of the test system, and the EMC measurement domain has benefitted from this as well. Generally speaking, any technique that can reduce the amount of human intervention can reduce errors caused by manually reading and recording measurement results. Automation also verifies and maintains the integrity of measurement settings to ensure repeatable results, and produces a verifiable test environment.

At the software level, the “must have” list of features required by EMC software is a long one, the most fundamental being the ability to completely collect, evaluate, and document RFI voltage, power, and field strength in accordance with current standards. However, the complexity of EMC measurements also makes it essential that software have two clearly-defined methods of operation. The first allows less-experienced users to obtain reliable, repeatable results using predefined, standards-compliant test routines, and the second allows “veterans” to specify custom values for every parameter in order to define their own test routines.

Virtually all commercial EMC software provides these capabilities to varying degrees, along with the ability to be updated as standards evolve. Today’s EMC software is typically based on Microsoft Windows, which makes it possible to create a familiar user environment that eliminates the need to navigate the nuances of proprietary software developed in-house.

### FFT BENEFITS AND CHALLENGES

One of the most recent and promising developments in EMC testing is the use of time-domain scanning methods based on the FFT technique to identify the disturbance spectrum. This approach has demonstrated its ability to reduce preview measurement time by a factor of 1000 or more. It is currently being evaluated by standards committees to determine whether it should be included in forthcoming modifications, but its viability has already been proven in a variety of measurement situations. To understand the benefits of this

<table>
<thead>
<tr>
<th>Frequency range</th>
<th>6-db bandwidth</th>
<th>Dwell time (s)</th>
<th>Minimum test time using analog measurement receiver</th>
</tr>
</thead>
<tbody>
<tr>
<td>30 Hz to 1 kHz</td>
<td>10 Hz</td>
<td>0.15</td>
<td>0.015 s/Hz</td>
</tr>
<tr>
<td>1 to 10 kHz</td>
<td>100 Hz</td>
<td>0.015</td>
<td>0.15 s/kHz</td>
</tr>
<tr>
<td>10 to 150 kHz</td>
<td>1 kHz</td>
<td>0.015</td>
<td>0.015 s/kHz</td>
</tr>
<tr>
<td>150 kHz to 30 MHz</td>
<td>10 kHz</td>
<td>0.015</td>
<td>1.5 s/MHz</td>
</tr>
<tr>
<td>30 MHz to 1 GHz</td>
<td>100 kHz</td>
<td>0.015</td>
<td>0.15 s/MHz</td>
</tr>
<tr>
<td>Above 1 GHz</td>
<td>1 MHz</td>
<td>0.015</td>
<td>15 s/MHz</td>
</tr>
</tbody>
</table>

**Table 1: Minimum CISPR 16 sweep times for peak and quasi-peak detection**
Time-Saving Effects of FFT-Based EMI Measurements

FEATURE

To reliably detect a pulse-like disturbance, the observation time per frequency point must be at least as large as the reciprocal of its pulse rate. In addition, disturbance measurements must always be made at the maximum level (e.g., the worst case emission), which usually requires repositioning the antenna and test device.

For example, scanning 30 MHz to 1 GHz with an IF bandwidth of 120 kHz and a step width of 40 kHz to measure the entire spectrum (without gaps and with sufficient measurement accuracy) produces 24,250 measurement points. If the dwell time is 10 ms per frequency point, total measurement time for a single preview scan is 4 min. This time must be multiplied by a factor of 20 or more to account for the time required for positioning the turntable and antenna height, and antenna polarization switching.

Using a spectrum analyzer instead of a test receiver does not overcome the problem because the time of a single sweep must be long enough for at least one disturbance pulse event to fall into the instrument’s resolution bandwidth at each frequency.

For repetitive sweeps and maximum hold for the trace display, observation time must continue until the spectrum becomes stable, and a continuous broadband signal will require many fast sweeps to show the envelope of the broadband spectrum. Spectrum analyzers usually allow fewer sweep points than test receivers and they may not provide enough frequency resolution to measure radiated emissions, which makes it necessary to perform time-consuming partial sweeping.

Conventional EMI measurement systems can only measure the signal within the resolution bandwidth within a stated measurement time, whereas FFT-based time-domain EMI measurement systems allow a much wider part of the observed spectrum to be analyzed simultaneously. This is because the EMI test receiver samples successive sections of spectrum at the IF with a bandwidth of several megahertz rather than only 120 kHz, and each “subspectrum” is calculated simultaneously with a specific resolution using FFT.

### Table 2: Bandwidth and measurement time specified by MIL-STD-461F

<table>
<thead>
<tr>
<th>Frequency band</th>
<th>Peak detection</th>
<th>Quasi-peak detection</th>
</tr>
</thead>
<tbody>
<tr>
<td>Band A: 9 kHz to 150 kHz</td>
<td>100 ms/kHz: 14.1 s</td>
<td>20 s/kHz: 2820 s (47 min.)</td>
</tr>
<tr>
<td>Band B: 150 kHz to 1 GHz</td>
<td>100 ms/kHz: 2985 s</td>
<td>200 s/MHz: 5970 s (1 h 38 min.)</td>
</tr>
<tr>
<td>Band C/D: 30 MHz to 1 GHz</td>
<td>1 ms/MHz: 0.97 s</td>
<td>20 s/MHz: 19,400 s (5 h 23 min.)</td>
</tr>
</tbody>
</table>
TIME-DOMAIN SCAN CONSIDERATIONS

However, steps must be taken when applying the time-domain technique that ensure all types of signals that can appear in a disturbance spectrum are correctly detected, even intermittent types with a very low pulse repetition frequency. If they are not considered, the frequency spectra calculated by the FFT may be displayed incorrectly in level and frequency.

Theoretically, an exact calculation of the frequency spectrum of a time-domain signal would require an infinite period of observation, and it would be necessary to know the signal amplitude at every point in time. In practice however, these requirements are unrealistic using FFT aided by digital signal processing. Analog-to-digital conversion provides a continuous input signal to be converted into an amplitude- and time-discrete signal, and applying the FFT limits signal observation time to a finite (and practical) amount. This means that calculation of the frequency spectra requires a reasonable number of discrete signals in the time domain, a process called “windowing.”

If the length of this window does not exactly correspond to an integer multiple of the periods of the frequencies contained in the input signal, it results in spreading or leakage of the spectral components away from the correct frequency and an undesirable modification of the total spectrum. The generation of spectral components that are not present in the original time-domain signal is known as the “leakage effect,” and is most severe when a simple rectangular window is used. The best way to reduce this effect is to choose a suitable window function that minimizes spreading.

The spectrum calculated by the FFT is a discrete frequency spectrum consisting of individual frequency components at the so-called “frequency bins,” which are determined by the FFT parameter. The original spectral response can only be observed at the discrete frequency bins, and there may be higher amplitudes in the original signal spectrum at frequencies between two adjacent frequency bins. The amplitude error this causes is called the “picket fence effect” (Figure 1) and is also characteristic of conventional stepped-frequency scans.

Time-domain measurement techniques employing FFT on intermittent disturbance signals require certain system parameters to be emphasized so that all disturbance signals are detected and measurement accuracy is maintained. For example, when an impulse-type disturbance signal is captured by the Gaussian-type FFT window, signal amplitude may be reduced at the window edges. To minimize this error while also ensuring that no signal is missed, EMI test receivers that employ time-domain scan include an overlap of the window function in the time domain.

Such receivers usually provide two settings for the step mode of the time-
domain scan, the “Auto CW” mode and an “Auto Pulse” mode. In Auto CW mode, the overlap in the time domain is about 20%, which allows narrowband signals to be analyzed as quickly as possible. The “Auto Pulse” mode provides more than 90% of overlap and is intended for broadband-impulsive and mixed signals. It ensures that even very short impulse signals at the edge of the Gaussian-type time-domain window are calculated without significant amplitude error. With so much window overlap, only a small amount of ripple remains in the time domain that could result in only a small measurement error.

Worst-case amplitude errors for such receivers are typically 0.4 dB for the lowest point of the amplitude ripple referred to the maximum pulse amplitude and the resulting average error is 0.09 dB, a theoretical value for a minimal pulse width. The real error value depends on the pulse duration and is usually even less. When performing the time-domain scan with weighting detectors to CISPR (i.e., quasi-peak), correct detection of single pulses requires the data rate for internal digital signal processing to be sufficiently high to accommodate the IF bandwidths that are used, and a 90% overlap of the FFT windows is essential for proper quasi-peak detection.

Analog filtering in the signal path has an influence on the frequency response of a time domain scan, and non-ideal correction of the analog filters in the RF and IF signal path of the test receiver add to overall measurement uncertainty. The bandwidths of the preselection filters get narrower as frequency decreases (such as 2 MHz bandwidth at 8 MHz vs. 80 MHz bandwidth at 500 MHz).

To minimize the influence of the preselection filters’ frequency response, the receiver reduces the bandwidth for the time-domain scan accordingly, from 7 MHz to 150 kHz for example, depending on the scan range, and compensates for the frequency response of the analog IF filters.

COMPARING STEPPED-FREQUENCY AND TIME-DOMAIN SCANS

In frequency bands A to E, the CISPR16-1-1 standard specifies bandwidths and tolerance masks for IF filters used in disturbance measurements to commercial standards. In

![Figure 3: Measured IF selectivity for CISPR bands A, B, and C/D](image)

![Figure 4: Overall frequency response of the test receiver for the time-domain scan (blue) and stepped-frequency scan (black) including the frequency response of the CISPR pulse generator](image)
contrast, MIL-STD-461 defines 6-dB bandwidths in decimal steps that must be met with a 10% tolerance. Any deviations from the specified tolerances cause amplitude errors.

To verify IF selectivity, a time-domain scan with max. peak detection was performed for sinusoidal test signals. A single measurement is insufficient for correct verification because the spacing of adjacent frequency bins is set to one-quarter IF bandwidth (Figure 2). The tests were repeatedly performed, increasing the start frequency of the time-domain scan step-by-step in small increments. All received frequency points were then merged into a single trace (Figure 3).

At lower levels, the inherent noise of the receiver limits the dynamic range and is specified as displayed average noise level (DANL). At higher levels, the nonlinearity of mixers and amplifiers limit the measurement range, and is characterized by the 1-dB compression or third-order intercept points. Sensitivity of 1 dB and 3 dB are analogous to these points, that is, where signal-to-noise ratio is high enough so that noise-induced measurement error is not more than 1 or 3 dB. Dynamic range usually specifies the usable level range between 1-dB sensitivity and the 1-dB compression point.

A measurement using a pulse generator for CISPR bands C and D compares the frequency responses of the stepped-frequency and time-domain scans (Figure 4). An exact evaluation of receiver measurement uncertainty is not possible with this measurement, and it does not consider errors caused by the cable and pulse generator, such as frequency response, matching, and long-term stability. However, it nonetheless shows that the differences between the two scan types are negligible.

Figures 5a, 5b, 5c and 5d: Comparison of resolution bandwidths with stepped-frequency scan (blue) and time-domain scan (green) for the CISPR 16 and MIL-STD 461 standards
Figures 5a through 5d show the measured frequency response of the CISPR bandwidths of 200 Hz, 9 kHz, and 120 kHz, and the MIL-STD IF bandwidths 100 kHz and 1 MHz for the time-domain scan and the stepped-frequency scan. Both traces match very well and are compliant with the requirements of the standards. Table 3 shows that the time-domain scan offers a higher dynamic range than the stepped-frequency scan, generally without regard to IF bandwidth.

The evaluation of measurement times was based on the frequency bands for EMI measurement in accordance with CISPR 25 (EN 55025) for automotive products and other military and commercial standards. CISPR 16-2 requires the measurements to be long enough so that at least one signal from the disturbance source is detected. For this comparison, the measurement time per frequency step for the commercial standards was set to 10 ms or 20 ms to correctly detect impulsive disturbances down to a pulse repetition rate of 100 Hz or 50 Hz respectively.

For measurements to MIL-STD-461, the measurement time was set according to Table 2. The measured values show that the time-domain technique considerably reduces the time required to perform a frequency scan even when using quasi-peak weighting and a dwell time of 1 s. The exact reduction depends on the IF bandwidth.

In short, FFT-based time-domain scan for preview measurements allows EMI testing to be performed in accordance with CISPR 16 orders of magnitude faster than when using a stepped-frequency scan. The measurement uncertainty of the time-domain and stepped-frequency scans is nearly identical. However, the stepped-frequency scan technique remains a proven, widely-accepted method, so it makes a great deal of sense to combine both the frequency and time-domain techniques throughout the design and certification process to ensure the best possible results.

**SUMMARY**

FFT-based time-domain scanning, along with the increasingly formidable capabilities of EMC software and greater process integration and automation, are transforming the EMC measurement process. As commercial and military standards evolve, these benefits will become more and more important, as will their ability to make the process easier for designers who already have a “full plate” of measurements necessary to bring a product to market.

Vic Hudson is product manager for EMI, OTA, and Antennas at Rohde & Schwarz, and can be reached at vic.hudson@rsa.rohde-schwarz.com.
Defining Who We Are

This is the first in a series of articles that, instead of focusing on technical aspects of EMC design, engineering, testing, and standards which all of us enjoy reading, I am taking a different approach as a contributing editor to examine in a controversial manner who we are and where our career may possibly take us in the future.

In reality we work as an electrical engineer, not an EMC engineer. The field of EMC is one aspect of electrical engineering. Not all products contain digital components or operate at frequencies above 1 MHz. Anything that uses electrical power, either AC or DC, involves the expertise of an engineer. What kind of engineer do you consider yourself (if involved in circuit design – are you an analog engineer or digital engineer)? What about the technology of today with circuits operating in the GHz range. Do you call yourself a digital engineer or a digital microwave engineer? When I ask someone how they classify themselves, they generally respond with either analog, digital or EMC. In reality, there is no such field as analog, digital, digital microwave, or microwave engineering….we are all in reality only one type of engineer-electrical.

The word digital is technically invalid when used in the field engineering. The word digital comes from the word digit and digitus (Latin for finger), as fingers are used for discrete counting [Wikipedia]. A digital engineer is in reality, “An infinitely fast AC slew rate signal engineer.” Since putting all these words on a business card is too long, we shorten it to digital. Draw a sine wave and then make the slopes really fast, with a period of time between rising and falling edges. It now looks like a digital pulse. Since there must be a finite time period for a waveform to go from 0V to voltage potential, we can never have a true digital signal, which implies an instantaneous edge transition. Conversely there is no such thing as a digital component. The input and output of every devices classified as digital is in reality an op-amp, which is an analog component. Technically, we are all analog engineers.

In future articles, we will examine technologies that are on the horizon and how the field of EMC will evolve into an even more exciting field.

Mark I. Montrose is an EMC consultant with Montrose Compliance Services, Inc. having 30 years of applied EMC experience. He currently sits on the Board of Directors of the IEEE (Division VI Director) and is a long term past member of the IEEE EMC Society Board of Directors as well as Champion and first President of the IEEE Product Safety Engineering Society. He provides professional consulting and training seminars worldwide and can be reached at mark@montrosecompliance.com

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INTRODUCTION

Everyone in the EMC business is familiar with the traditional Normalized Site Attenuation test (NSA). However, in February of 2007 CISPR 16-1-4 was published complete with the new Site Voltage Standing Wave Ratio (SVSWR) test. At the time, the American National Standards Institute (ANSI) Accredited Standards Committee (ASC) C63® had developed a draft proposal for C63.4 (Draft 1 - May 20, 2005) called the Time Domain Reflectivity (TDR) measurement. The critical question addressed by this article is which method – SVSWR or TDR - more accurately provides an assessment of the test site. Given the investments companies make in test sites for EMC compatibility, this is key assessment question.

BACKGROUND

Conceptually, the SVSWR method is quite straightforward and easily understood. As with any VSWR measurement the objective is to measure the maximum and minimum values of a standing wave as illustrated in Figure 1. The ratio of these values is the VSWR. The most common application of the VSWR measurement is in evaluating transmission lines. If there is an impedance mismatch at the end of a transmission line between the impedances of the transmission line and the load (for example), there will be a boundary condition that results in a reflected wave. The reflected wave will, at various locations on the transmission line, be constructively or destructively interacting with the continuous wave from the source. The resulting construct (direct and reflected wave combination) is a standing wave. A simple example of this is found in
the conducted power test required for appliances in CISPR 14-1. In this test a transducer (power clamp) is moved along an extended power cord of the product in an effort to measure the maximum voltage on the power cord over the frequency range of interest. The same event is realized on an imperfect test site. The transmission line is the path from the equipment under test to the receiving antenna. Reflected waves are created from other objects in the test environment. Those objects could range from chamber walls to buildings and cars (at open area test sites). Just as in the case of a transmission line, a standing wave is created. The test set up for the site VSWR or SVSWR test is shown in Figure 2.

**UNDER SAMPLING**

The physical dimensions of the standing wave are a critical factor in accurately measuring a standing wave. The objective, again, is to find the maximum and minimum value. The SVSWR test in CISPR 16-1-4 proposes to measure the standing wave on a test site by moving a transmitting antenna along a straight line in the chamber and measuring the received voltage with the emissions antenna in the normal location used for product testing. Just as in a conducted power test or similar VSWR measurement, a continuous movement of the transducer, or in the case of SVSWR the transmitting antenna, is needed to ensure the capture of the maxima and minima of the standing wave. This could be done at each frequency but only at considerable expense and time. Consequently, the CISPR working group decided to compromise and measure only six physical positions for each of the volumetric locations (see Figure 3 on page 46). The only other option for reducing the test time was to reduce the frequency resolution of the measurement (e.g. measure fewer frequencies but at each frequency measure more positions). The problem with that option is that many objects that reflect can have narrow spectral characteristics. In other words, some materials can be significantly reflective for a narrow frequency range. Consequently, the working group decided to apply a maximum 50 MHz step size to the test resulting in a minimum of 340 frequencies from 1-18 GHz but with only six positions as shown in Figure 3 (page 46).

The sampling of a standing wave at only a discrete number of positions may plausibly provide sufficient accuracy to compute an approximate SVSWR depending on the size of the steps. However, another compromise was to have the same prescribed positions for every frequency so that the test would save time by moving the antenna and sweeping frequency. The chosen positions are 0, +2, +10, +18, +30, +40 cm. Try to imagine a sign wave superimposed on a ruler with six marks on it. Now imagine compressing the sign wave into shorter and shorter wavelengths. Figure 4 (page 46) illustrates this thought experiment. There will be frequencies where the chosen locations will never come close to the true maxima or minima of the sign wave. This is a compromise that will result in a compliance bias, e.g. a result that is always lower than the true SVSWR. This bias is an error term and should not be confused with a measurement uncertainty contribution.

How large is the error term? If we think of the example illustrated in Figure 4 (page 46) it is clear the wavelength is 2 centimeters. That would be a 15 GHz sign wave. At that frequency, there would be no measured standing wave because the wavelength is 2 cm and the other locations are even multiples of 2 (10, 18, 30 and 40 cm)! Of course, the same issue occurs at 7.5 GHz. At virtually every frequency the sampling results in measuring neither the maximum nor the minimum.

**TEST TIME**

A laboratory must measure four locations as shown in Figure 3 (page 46) in two polarities and at least two heights in accordance with CISPR 16-1-4. The measurement range is 1-18 GHz. Until recently, the only antennas available that met...
the pattern requirements were available in 1-6 GHz and 6-18 GHz models. The consequence is that the test time is shown in Equation 1:

\[
\text{Time} = \text{t}_{\text{set up}} + \left[ (\text{t}_{\text{move antenna}} + \text{t}_{\text{sweep}}) \times n_{\text{positions}} \times \right. \\
\left. n_{\text{locations}} \times n_{\text{heights}} \times n_{\text{antennas}} \times n_{\text{polarities}} \right] + \\
(n_{\text{antennas}} \times n_{\text{switch antennas}}) + \left( n_{\text{heights}} \times n_{\text{change heights}} \right) + \left( n_{\text{polarities}} \times n_{\text{change polarities}} \right)
\]

Where: \( t_x \) = time to perform function \( x \), \( n_y \) = number of times activity \( Y \) must be performed.

**Equation 1: Estimate test time for SVSWR**

The result of this combination of positions, locations, polarities, heights and antennas results in a rather lengthy test. This time represents an opportunity cost to the laboratory. The opportunity cost is the revenue that could otherwise have been realized in lieu of conducting this lengthy test. As an example, a typical test time for this test is at least three test shifts. If a lab were to charge $2,000 USD for a shift, this test represents an annual opportunity cost, assuming the site is checked annually as recommended, of at least $6,000-$12,000 USD. This does not include the initial costs of the special antennas ($14,000 USD).

**POSITIONING UNCERTAINTY**

Each measurement of the SVSWR method requires the positioning of the transmitting antenna to the positions specified (0, 2, 10, 18, 30, 40 cm). Since the computations are corrected for distance, the repeatability and reproducibility of the positioning directly impacts the measurement uncertainty. The question then becomes, how repeatable and reproducible is the positioning of the antennas in increments as small as 2 cm? A recent gage study conducted at UL has demonstrated this contribution to be approximately 2.5 mm or about 15% of the 18 GHz wavelength. The magnitude of this contributor will depend on frequency and the amplitude of the standing wave (an unknown).

A second factor related to positioning is angle versus the antenna pattern. The antenna pattern requirements in CISPR 16-4-1 has variability of roughly +/-2 or 3 dB in H-plane and even wider in E-plane. If you pick two antennas with different patterns but both meet the pattern requirements, you can have very different results. In addition to this antenna to antenna variability (a reproducibility problem), the antennas used to transmit do not have perfectly symmetric patterns (e.g. patterns vary with small increments in angle) as shown in the standard. As a consequence, any change in alignment of the transmitting antenna to the receiving antenna results in a changed received voltage (a repeatability problem). Figure 5 (page 48) illustrates the actual pattern changes of a SVSWR antenna with small increments in the angle. These true pattern characteristics result in significant angular positioning variability.

The changes in antenna gain as a function of relatively small angular rotations causes as much as 1 dB of variability in the example shown.

**TIME DOMAIN METHOD TO OBTAIN SVSWR**

The SVSWR method in CISPR 16-1-4 is based on moving antennas spatially to vary the phase relationship between the direct wave and reflected waves from chamber imperfections. As discussed previously, when the waves add constructively, there is a peak response \( E_{\text{max}} \) between the two antennas and when the waves add destructively, there is a minimum response \( E_{\text{min}} \). The transmission can be expressed as

\[
E = E_D + \sum_i^N E_R(i)
\]

where \( E \) is the received field strength.

\( E_D \) is the direct path signal, \( N \) is total number of reflections from the site (this could include single or multiple reflections from the chamber walls or open area site imperfections). \( E_R(i) \) is the \( i^{th} \) reflected signal. For ease of the derivation, let us assume there is only one reflected signal (this will not lose the generality). The site VSWR (or the relative ripple size) of the site can be expressed as

\[
S = \frac{E_{\text{max}}}{E_{\text{min}}} = \frac{E_D + E_R}{E_D - E_R} = \frac{1 + E_R/E_D}{1 - E_R/E_D}
\]
The Application of Time Domain Measurements at Northwest EMC

Greg Kiemel, Northwest EMC, Inc.

Several years ago, we decided to purchase a 20 GHz Vector Network Analyzer (VNA) to improve the quality of our calibrations and to perform site validation measurements. Previously, we had been using a 500 MHz VNA to calibrate LISNs, CDNs, and current probes; but we only had a basic understanding of network analysis. We didn’t really grasp the benefits of time-domain reflectometry (TDR) that are available in modern VNAs.

It wasn’t until we witnessed a TDR demonstration by Mike Windler of UL, that we began to understand the possibilities. Mike performed site validation above 1 GHz using a proposed revision of ANSI C63.4 (circa April 2005) in one of our 10m chambers. The benefits were immediately obvious: the TDR method was much faster than the Site Voltage Standing Wave Ratio (SVSWR) method, it was much less labor intensive, and troubleshooting with the TDR method was far superior. In fact, it appeared to be the only way to identify the source of non-compliance.

Despite the many benefits of using the TDR method for site validation, as an accredited test laboratory we are compelled to also perform site validation measurements using the CISPR 16-1-4 SVSWR method. The SVSWR method is also cited in the latest edition of CISPR 22 and the EuroNorm version (EN55022) will be a requirement in Europe and Japan next year (note at the time of this writing there may be some delay). We decided to conduct a study where both methods were used on a variety of test sites (10m, 5m, and 3m chambers, and a 10m OATS). Our hope was to correlate the results, so we would use the TDR method to not only troubleshoot problems with our test sites, but also confirm their continued compliance.

The study was conclusive in many aspects:

- Both TDR and SVSWR methods correlate extremely well in determining compliance
- Our test facilities which were measured with the SVSWR and the TDR techniques passed the respective site validation requirements. This was considered very useful in accepting the TDR technique as well as the SVSWR technique.
- RF absorber type, the coverage area of the absorber, and the chamber volume are all factors in meeting site validation requirements
- The SVSWR method is more labor intensive and utilizes more of the existing lab equipment
- The TDR method is an excellent tool in identifying the source of non-compliance and is much faster

So in addition to using our VNA to perform equipment calibrations, it is used on an on-going basis to improve the performance of our test sites. The TDR function is terrific in identifying the exact fault location in cables and fixtures. It is also the best tool to measure absorber performance.

We also make the VNA available to our clients as a product development tool. It is extremely well suited to evaluate antenna matching networks as well as antenna performance. As a result, we’ve been able to grow our business while lowering our cost for calibrations and site verifications.

Greg Kiemel is the Director of Engineering at Northwest EMC, Inc. He has 23 years experience in the EMC field. Mr. Kiemel is a NARTE-certified EMC and ESD engineer, as well as a Senior Member of the IEEE. He is active in the ANSI ASC C63® and TCBC committees. Mr. Kiemel recently completed his tenure as a Distinguished Lecturer for the IEEE EMC Society. Prior to his fifteen years with Northwest EMC, he worked as the lead regulatory engineer in the personal computer division at Epson Portland, Inc. and as an EMC engineer at Tektronix, Inc. He earned his BS in Engineering from Weber State University. Mr. Kiemel may be reached via email at gkiemel@mwemc.com or phone at 888-364-2378.
By solving Equation 3, we obtain the ratio of the reflected signal to the direct signal

\[ E_{\text{relative}} = \frac{E_R}{E_D} = \frac{1-S}{1+S} \]  

(4)

As can be seen from Equation 4, the two terms, i.e. the reflected to direct signal ratio \( E_{\text{relative}} \) and the site VSWR \( S \) describe the same physical quantity – a measure of the level of reflections in the site. By measuring the site VSWR (as is the case in CISPR 16-1-4), we can determine how large the reflected waves are relative to the direct wave. In an ideal situation there is no reflections, resulting in \( E_{\text{relative}} = 0 \), and \( S = 1 \).

As previously discussed, to detect the ratio between the reflected and the direct signal, in the site VSWR method in CISPR 16-1-4, we change the separation distance so the phase relationship between the direct path and reflected signals can be varied. Subsequently, we derive the SVSWR from these scalar responses. It turns out that we can acquire the same SVSWR using vector (voltage and phase) measurements without the need to physically move the antennas. This can be done with the aid of a modern vector network analyzer (VNA) and time domain transformations. Notice that Equations 2 to 4 hold true in either frequency domain or time domain. In time domain, however, we can distinguish the reflected signals from the direct signal because the point in time at which they arrive at the receive antenna is different. This can be viewed as a pulse sent out from the transmit antenna. In time domain, the direct wave will arrive at the receive antenna first, and the reflected wave will arrive later. By applying time gating (a time filter), the effect of the direct signal can be separated from the reflected ones.

The actual measurements are performed in frequency domain with a VNA. The results are then transformed to time domain using inverse Fourier transform. In time domain, time gating is applied to parse the direct and reflected signals. Figure 6 shows an example of the time domain response between two antennas (by using inverse Fourier transform from frequency domain measurements). Figure 7 shows the same time domain response with the direct signal gated out. The time domain data (after the parsing) are finally converted back to frequency domain using Fourier transform. For example, when the data in Figure 7 is transformed back to frequency domain, it represents \( E_R \) versus frequency. In the end, we obtain the same \( E_{\text{relative}} \) as the CISPR spatial varying method,
Moving Forward in ANSI ASC C63® and IEC/CISPR with Time Domain Measurements

Don Heirman, Don HEIRMAN Consultants

The 2003 and 2009 editions of ANSI C63.4 have recently been recognized by the FCC in a Public Notice issued in November for compliance measurements for product certification. In the 2009 edition, there are two methods for site validation above 1 GHz. One is to have absorber laid down in a particular pattern between the place where the EUT is placed and the measurement antenna location. The absorbers have to have a specific performance identified in the 2009 edition. The second method in making site validation measurements is to use the method in IEC/CISPR in their publication 16-1-4. This is referred to as the site VSWR method (S-VSWR). This is based on a series of measurements at the extremes of the EUT volume occupied on the test site using a specified transmit and receiving antennas aimed at each other and test equipment owned by every test laboratory. This work was based on years of effort that included practical experimentation by several test labs. Presently test site validation is performed using this technique internationally.

In ASC C63®, work has been proceeding in another method that has much promise as indicated in this article. This is part of the work that will lead to the publication C63.25. It has been shown that using this time domain approach will add value in that it not only will determine if a site meets validation requirement but also locate areas in the test site where the site may need to be improved or rid itself of a reflective source to meet the site validation requirement using this technique. To facilitate this test, a vector network analyzer is needed which may not be available to a test lab but can be rented for the purpose. Experimentation with this technique has been ongoing with test labs showing its usefulness. Preliminary results that both the S-VSWR and time domain techniques have given similar site validation results, i.e. sites meet the acceptance criteria for both validation techniques. This is quite helpful as the test lab does not want to perform two validation techniques to show site acceptance. Once there is experience with the time domain technique, it should be considered for introduction into C63.4 or referenced as one of the options for site validation above 1 GHz.

CISPR has been introduced to time domain concept in the past. The work on the S-VSWR received the immediate attention and hence was published. ASC C63® is encouraged to suggest time domain site validation techniques to CISPR as a US contribution. The best time for this is once C63.25 is published; it is always preferable to base inputs to CISPR on published standards that are used by industry. This will show the usefulness and practical application for test labs. Such usefulness is also a goal of CISPR as it provides basic standards covering measurement methods and instrumentation (including test sites). Hence, the users of C63.25 when it is published will be encouraged to bring the matter to the attention of members of the technical advisory committee of the US National Committee of the IEC/CISPR. All inputs to the CISPR must come from a member of the working group for in this case, CISPR Subcommittee A Working Group 1. While this concept has been discussed, it is time again to revisit it with the CISPR working group by tabling a document for consideration at the WG’s next meeting in Seattle in October 2010.

This then leads to a call for help in working with the new technique and the drafting of C63.25 as well as introducing it into the CISPR working group noted above. Please contact Don Heirman at d.heirman@att.net if there are questions on your willingness to work this exciting project(s).

Donald Heirman is president of Don HEIRMAN Consultants, training, standards, and educational electromagnetic compatibility (EMC) consultation corporation. Previously he was with Bell Laboratories for over 30 years in many EMC roles including Manager of Lucent Technologies (Bell Labs) Global Product Compliance Laboratory, which he founded, and where he was in charge of the Corporation’s major EMC and regulatory test facility and its participation in ANSI accredited standards and international EMC standardization committees. He chairs, or is a principal technical contributor to, US and international EMC standards organizations including ANSI ASC C63® (chairman) and the International Electrotechnical Commission’s (IEC) Special International Committee on Radio Interference (CISPR) where in October 2007 he was named the chair of CISPR moving from his previous role as its subcommittee A chairman responsible for CISPR Publication 16 on basic EMC measurement methods, test instrumentation requirements and statistical methods. He is a member of the IEC’s Advisory Committee on EMC (ACEC) and the Technical Management Committee of the US National Committee of the IEC. In November 2008 he was presented with the prestigious IEC Lord Kelvin award at the IEC General Meeting in Sao Paulo, Brazil. He is a life Fellow of the IEEE and a life member of the IEEE EMC Society (EMCS) and member of its Board of Directors, chair of its technical committee on EMC measurements, past EMCS president and vice president for standards, and past chair of its standards development committee. He also is past president of the IEEE Standards Association and past member of the IEEE Board of Directors.
FEATURE Site Voltage Standing Wave Ratio versus Time Domain Reflectometry

but by going through a different route. Although the inverse Fourier transform (or the subsequent Fourier transform) sounds like a daunting task, it is actually a built-in function in a modern VNA. It takes no more than the pushing of a few buttons.

NEXT STEPS: IMPROVING THE TIME DOMAIN SVSWR METHOD FURTHER

We have established that the SVSWR by spatial movement and SVSWR by time domain produce equivalent data. Empirical measurements can validate this point. Questions that still linger are: whether this is the most representative data for Equipment Under Test (EUT), and what uncertainties we can achieve due to antenna selections? Referring to Equation 2, all reflections are modified by the antenna pattern before being summed. For simplicity, let us consider a test chamber where multi-reflections are negligible. We then have seven terms in the transmission path, namely the direct signal, and reflections from four walls, the ceiling and the floor. In CISPR 16-1-4, there are very specific requirements on the transmitting antenna pattern. For practical reasons, these requirements are by no means restrictive. For example, assume the back wall reflection is the dominant imperfection, and the front to back ratio of the antenna is 6 dB (within CISPR 16 specification). For a site with a measured SVSWR=2 (6 dB) using a perfect isotropic antenna, \( E_s/E_D \) is 1/3. If we use an antenna with a front-to-back ratio of 6 dB, the measured SVSWR becomes

\[
S = \frac{E_{\text{max}}}{E_{\text{min}}} = \frac{1 + 0.5 E_{\text{backwall}}}{1 - 0.5 E_{\text{backwall}}} / E_D = \frac{1 + 0.5 \times 1/3}{1 - 0.5 \times 1/3} = 1.4
\]

The antenna with a front-to-back ratio of 6 dB underestimates the SVSWR by 20*log(2.0/1.4) = 2.9 dB. The above example is obviously overly-simplified. When considering all other reflections of the chamber, and all variations of the antenna patterns, the potential uncertainty is even larger. In the other polarization (in E-plane), it is not possible to have a physical isotropic antenna. It is an even greater challenge to define a strict antenna pattern, which all real physical antennas must meet.

The quandary related to pattern variations can be solved by rotating the transmitting antenna. In this scheme, we do not need an antenna with a broad beam – a familiar double ridged waveguide antenna commonly used in this frequency range will work fine. It is still preferred to have a large front to back ratio (which can be easily improved by placing a small piece of absorber behind the antenna). The implementation is the same as discussed earlier for the time domain method, except that we also rotate the transmitting antenna by 360° and perform a maximum hold. Instead of trying to illuminate all walls at the same time, this scheme does it one at a time. This method may yield results that are slightly different from ATTEMPTING to broadcast to all walls at the same time. It can be argued that it is a better metric of a site performance, as a real EUT is likely to have a narrow beam rather than looking like a specifically crafted antenna. In addition to

MORE INFO

The term free-space implies that there is no electromagnetic interaction between the test environment and the antenna. Use of the time domain to separate spacial effects allows for determination of the environment or the antenna without influences of the other. A recent paper on this topic was presented, and selected as best symposium paper, at the 2007 IEEE International Symposium on EMC. The paper [1], titled “Free Space Antenna Factors through the use of Time-Domain Signal Processing” by Dennis Camell, Robert Johnk, David Novotny and Chriss Grosvenor of the National Institute of Standards and Technology (NIST), describes a process to determine free space antenna factors using the standard site method (SSM) without the accompanying facility effects. Time domain gating routines, usually built into a vector network analyzer (VNA), can be used to remove the reflected signals of the facility thus providing a free-space environment for the antenna. This process provides excellent results above 1 GHz and good results for some cases below. This method allows for improved accuracy in the determination of free-space antenna factors. Finally, this method fits well with current EMC standards methods.

ANSI ASC C63®, a US national standards committee on EMC, has working groups that are leading efforts to include time domain measurement methodology in EMC standards. This includes both the antenna area with the revision of the ANSI C63.5 standard and in site acceptability with a new standard, C63.25. These working groups are always looking for new members to help in this work.

Visit www.c63.org or contact Don Heirman at d.heirman@ieee.org for more information. Better yet, attend the next series of ANSI C63 meetings which will be held in New Brunswick, NJ at IEEE Headquarters the week of April 19, 2010. Contact the Subcommittee 1 Secretary, Janet O’Neil, for meeting information at j.n.oneil@ieee.org.

avoiding the messy situation due to the antenna patterns, we can pinpoint where an imperfection occurs in a chamber or an OATS. The location can be identified from the rotation angle, and time needed for the signal to travel (thus the distance to where the reflection occurs).

**CONCLUSION**

The benefits of the time domain method are numerous. It avoids the pitfall of the under-sampling issue discussed earlier. The method does not depend on physically moving the antennas to a few discrete locations, and the SVSWR from time domain represents the true value of the site. Also, in the CISPR method, to normalize the influence due to the path length, the exact distance between the antennas must be known. Any uncertainties due to the distance translate into uncertainties of the SVSWR (considering the small increments needed, it is even more challenging). In time domain, there are no distance normalization uncertainties. In addition, perhaps the most attractive feature for an end user is that time domain SVSWR is much less time consuming. The test time is reduced almost six fold (see Equation 1).

One might be tempted to argue that in the CISPR method, because the antennas are moved, the reflection points move on the chamber walls, and more areas of the imperfections are covered. This is a red herring. The purpose of moving the receive antenna is to vary the phase relations only. The total distance varied is 40 cm. It translates to 20 cm (7.9”) coverage on the wall due to geometry translations (if the transmission path is parallel to the chamber wall). For the theory to work out, we in fact need to assume the reflection properties of the absorbers are uniform along the whole 20 cm. To cover more areas, one needs to move the antennas much more drastically, as is done in CISPR 16-1-4 (the front, center, left and right locations).

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A fully anechoic chamber features absorber treatment on all four walls, floor and ceiling of the chamber. Time Domain Reflectivity (TDR) measurements not only can provide an accurate assessment of a test site such as this, but can also provide additional information such as where the largest contributors to deviations from an ideal site come from.
This paper describes how to remove the measurement artifacts caused by discontinuities in high frequency S-parameter data caused by the test connectors on the Printed Circuit Boards (PCBs) and cables. The frequency domain S-parameters are converted to the time domain to get the impulse response. Time domain gating is then used on this impulse response to remove reflections due to end connectors and/or other discontinuities. The gated impulse response is then transformed back to the frequency domain. The final result is a much improved S-parameter data set with unwanted resonance removed, allowing the PCB trace or cable loss to be determined.
INTRODUCTION

High speed data signals are significantly affected by the losses associated with printed circuit board (PCB) and cable conductor and dielectric material. Making direct measurements of this loss using a Vector Network Analyzer (VNA) is a common method to determine if the PCB trace and/or cable will cause an unacceptable amount of loss for the desired data rate and path length. However, the lengths associated with most PCB traces, and certainly with most cables, result in the likelihood of reflections and resonance effects in the S-parameter data when the cable or trace is not perfectly terminated in the characteristic impedance of the cable or trace. These resonances are length dependent, therefore the S-parameter results do not accurately indicate the true effects of the trace/cable, but rather indicate the loss associated with a particular length of trace/cable, and with the specific connectors used. Due to these measurement artifacts, the measurement results are often not useful for more general analysis of PCB traces (without connectors, but directly connected to an IC), or with cables that may use different connectors, or when the length of the trace/cable may vary. It is usually desired to have a measure of the trace/cable loss per unit length (per inch, meter, etc) so that the S-parameters for any required length can be created from the original measured data.

One option to remove unwanted connector effects is to use TRL (thru, reflect, line) VNA calibration. This VNA calibration method effectively removes these end effects by performing the calibration right on the PCB. However, this technique requires a number of different length traces on a PCB in order to calibrate over a wide frequency range. Also, the connectors themselves must be high quality connectors so their high frequency characteristics are all similar, as well as the dielectric material on the PCB must be consistent across the PCB. These requirements are not always possible, especially when using PCBs with low cost dielectric, and TRL does not apply for cables.

This paper shows a number of examples of the effects of time gating on PCB traces and cables. The agreement is very good, indicating that this procedure effectively removes the unwanted reflections/resonances and allows an accurate determination of the per-unit-length parameters for traces and cables.
TIME GATING

Time domain gating is simply applying a windowing shape [1] to the time domain data, effectively weighting the time domain data and reducing the undesirable effects. There are many different windowing functions available: Rectangular, Hann1, Hamming, Blackman, Kaiser, and Cosine Squared are only a few of the more popular window functions. Each windowing function has an associated shape and is more or less complex to calculate. In general, the more complex (time consuming) that a given windowing function becomes, the better it reduces side lobes in the frequency domain. Figure 1 shows a few examples of the more popular window functions.

By applying time domain gating (windowing) of frequency domain data, the effects of end connectors, etc. can be removed. The original frequency domain data (magnitude and phase) are transformed using an Inverse Fourier Transform to give a time domain impulse response. The time domain impulse response is then gated using a modified Hann window2. The Hann window was modified to use the first half of the Hann window in Figure 1, but delayed until the start of the time domain pulse. This allows the high frequency portion of the impulse response to be maintained, while smoothing the late-time effects of the measurement artifacts.

The overall length of the gating window can be adjusted to remove unwanted end connector reflections. Once the gated impulse response is complete, the data are transformed back to the frequency domain. The length of the gating window determines the lowest valid frequency in the gated frequency domain data, and a low frequency correction using the original low frequency data, is required as the final step. This final step is important or the results below the lowest valid frequency will be incorrect. The resulting S-parameter data allow for a straightforward scaling to the desired length, since the end connector reflections/resonances have been removed.

Figure 2 shows an example of a PCB S21 measurement using SOLT calibration (ungated). Resonances cause obvious peaks and valleys in the S21 measurement above 2 GHz. These peaks and valleys can cause false signal integrity simulation results if the fundamental or harmonic frequencies of the signal fall into a peak or valley.

The ungated frequency domain signal was converted to the time domain using an inverse Fourier Transform. Figure 3 shows the resulting impulse response from the ungated signal in Figure 2.

After the main pulse in Figure 3, there are a number of reflections that can be seen. However, since Figure 3 is using a linear vertical scale, these reflections do not seem important.

Since the scale in Figure 2 is a log scale, and the frequency range of interest has results as low as -60 dB, the time domain signal in Figure 3 is expanded in Figure 4. It is clear in Figure 4 that significant reflections occur.

When a modified Hann window is used the time domain pulse is without significant reflections. The rectangular front edge

---

1 The Hann window is often called the “Hanning” window.
2 The Hann window was selected due to its simplicity and good performance.
The gating operation removes low frequency information. This example used a 1 ns gate width; therefore, the gate is not wide enough to include the correct frequency domain data below 1 GHz. Low frequency S21 data is usually accurate, so the gated impulse is corrected by replacing the initial gated results with the original low frequency results (below 1 GHz in this example). The final gated-corrected impulse is also shown in Figure 5.

The impulse response is converted back to the frequency domain, and the final gated-corrected S21 data is shown in Figure 2. The gated-corrected data clearly has improved the S21 curve by eliminating the peaks and valleys in the original data.

A second example uses measured S-parameter data for a five meter long cable. Figure 6 shows the original S21 magnitude data. The effect of length related resonances/reflections is apparent below 1 GHz, and dominates the values above about 3 GHz.

The time domain original data, gated data, and corrected-gated data are shown in Figure 7. In this case, significant reflections occurred before the main impulse, as well as after it. The early reflections were due to a poor connection between the network analyzer and the differential cable. Due to the construction of the cable, it was impossible to improve this connection. This shows another important use for this time gating technique. If this ‘early time’ data is included in simulations, the results would be non-causal and incorrect. The final gated-corrected impulse is much cleaner than the original data. All the early reflection noise is eliminated, and nearly all the later reflection noise as well.

The final frequency domain results are shown in Figure 8. The gate width was 3 ns, so the original frequency domain data below 333 MHz was used to correct the gated data at low frequencies. If Figure 6 and Figure 8 are compared, it is
clear that the original data has been significantly improved by eliminating undesired resonance effects.

**PHASE ERRORS IN MEASUREMENTS**

In the course of making the measurements with the VNA we happened to measure the same cable with both a high frequency VNA (upper frequency at 20 GHz) and a low frequency VNA (upper frequency at 8 GHz). When the time domain impulse response was compared, the delay was different for the two different VNA measurements by a significant difference (Figure 9). The delay measured with the TDR was 42.8 ns matching the low frequency VNA. However, the high frequency VNA was reporting a much shorter delay for the same cable.

It was found that the high frequency VNA was set for 50 MHz steps. Figure 10 shows the original phase data before gating. The individual data points show that most of the phase wrapping for this cable was missed by the large frequency step size. This resulted in an incorrect and much lower phase delay (and hence less delay in the time domain) for the same cable. Once the frequency step size for the high frequency VNA was made small enough to capture the correct phase, the delay of the cable was the same for both VNAs.

The final S21 magnitude once the gating and low frequency correction was performed is shown in Figure 11. The gated data without the low frequency correction is also shown.

**SUMMARY**

While good measurement techniques and accurate calibration are extremely important, effects of reflections from imperfect signal launch (due to poor connectors and/or length resonances) can sometimes corrupt the desired data. This is especially true at high frequencies, at which the DUT becomes electrically long.

Even with careful calibration, the limitations of both the SOLT and TRL network analyzer calibration techniques often result in S-parameter data that is likely to give inaccurate results to signal integrity simulations.

The frequency domain S21 data were transformed into the time domain, and a modified Hann window was applied to data from both a PCB and a cable. Once the windowing was performed, the data were transformed back into the frequency domain.

The final S21 data were corrected for low frequencies by using the original low frequency S21 data.
The final S21 frequency domain data was clearly improved over the original data. End connector effects, and even a poor connection between the network analyzer, were eliminated using this gating procedure. Furthermore, care must be taken that the frequency step size is small enough so that phase information is not lost for long cables. This is usually not an issue for traces on a PCB since these traces tend to be short. Regardless of the path being measured, the issue of possible phase error in measurements must be considered and eliminated.

While not all network analyzer measurements require correction using gating, this procedure can extend the useful frequency range of simple SOLT calibration to higher frequencies than might otherwise be useful.

**REFERENCES**


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