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EMC DIGEST

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Top Three EMI and Power Integrity Problems with On-Board DC-DC Converters and LDO Regulators
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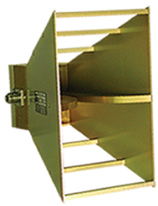

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FROM THE EDITOR:
INTRODUCTION TO THE EMC DIGEST

Kenneth Wyatt
Senior Technical Editor / Interference Technology / ITEM Media
E: kwyatt@interferencetechnology.com | T: (US) 719-310-5418

Welcome to this first edition of the EMC Digest, a compilation of some of Interference Technology's most popular articles during the last three years as chosen by our readers! As you can see, the subjects readers deemed most useful are related to EMC fundamentals, product design, and test.

As a product designer myself (too many years ago), a long time EMC engineer for HP and Agilent Technologies (now Keysight), and currently a consultant to hundreds of companies over the last ten years, I understand very well, the issues product designers face when it comes to EMC and EMI issues.

As companies worldwide have consolidated, merged, and generally downsized, the days of having a dedicated EMC engineer on staff for many companies has literally disappeared. Now product designers are being asked to wear many hats. Product compliance for EMC - is just one more of many. Unfortunately, this subject is rarely taught in colleges or universities.

During my own consulting over many years, I find most manufacturers continue to leave EMC compliance to the end of development cycles with the resulting panic as their "baby" fails various EMC compliance testing. Whenever my phone rings, oftentimes it's a project manager with a subtle, but underlying, panic in their voices calling for a quick resolution please! Sometimes we can offer a quick fix, but too many times the only fix is a redesign.

Despite all of us dozens of consultants worldwide who endeavor to train those designers "in the trenches" to deal with product compliance earlier in the design cycle, many simply aren't getting the word. And the issues always seem to revolve around the same old list of design issues: poor circuit board layout, inadequate shielding, lack of filtering, cables penetrating shielded enclosures, and other simple design features that could have easily been incorporated more cheaply into the design early on.

That's one reason my coauthor, Patrick André and I decided to write our best selling book, *EMI Troubleshooting Cookbook for Product Designers*, after commiserating with each other during an industry EMC conference on why we tend to deal with the same old issues time and again for our clients!

Enter this year's EMC Digest! You'll find articles on basic EMC concepts in order to lay out a solid foundation, the essentials of designing products for EMC compliance, designing EMC filters, an article on issues with DC-DC converters, and radiated emission measurements at various test distances.

For you product designers that end up dealing with EMC at the end of the design cycle, we include "EMI Troubleshooting - Step-By-Step" and a new emerging troubleshooting tool; "EMI Troubleshooting with Real-Time Spectrum Analyzers" - a real time-saver when it comes to debugging transient or infrequent EMI signals.

While all the above articles apply to military and aerospace applications, we also offer a summary of up to date military and aerospace EMC test standards.

We, at Interference Technology Magazine and ITEM Media, hope this new EMC Digest will find a home on your desk and that the contents will help you achieve success in your current and future designs! Best wishes!

BASIC EMC CONCEPTS

Kenneth Wyatt
Wyatt Technical Services LLC
ken@emc-seminars.com

Understanding EMC is all about two important concepts: (1) all currents flow in loops and (2) high frequency signals are propagated as electromagnetic waves in transmission lines.

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Currents Flow In Loops

These two concepts are closely related and coupled to one another. The problem we circuit designers miss is defining the return path back to the source. If you think about it, we don't even draw these return paths on the schematic diagram - just showing it as a series of various "ground" symbols.

So what is "high frequency"? Basically, anything higher than 50 to 100 kHz. For frequencies less than this, the return current will tend to follow the shortest path back to the source (path of least resistance). For frequencies above this, the return current tends to follow directly under the signal trace and back to the source (path of least impedance).

Where some board designs go wrong is when high *dV/dt* return signals, such as those from low frequency DC-DC switch mode converters or high *di/dt* return signals get comingled with I/O circuit return currents or sensitive analog return currents. We'll discuss PC board design in the next article. Just be aware of the importance of designing defined signal and power supply return paths. That's why the use of solid return planes under high frequency signals and then segregating digital, power, and analog circuitry on your board is so important.

How Signals Move

At frequencies greater than about 50 to 100 kHz, digital signals start to propagate as electromagnetic waves in transmission lines. As shown in *Figure 1*, a high frequency signal propagates along a transmission line (circuit trace over return plane, for example), and the wave front induces a conduction current in the copper trace and back along the return plane. Of course, this conduction current cannot flow through the PC board dielectric, but the charge at the wave front repels a like charge on the return plane, which "appears" as if current is flowing. This is the same principle for capacitors and Maxwell called this effect "displacement current".

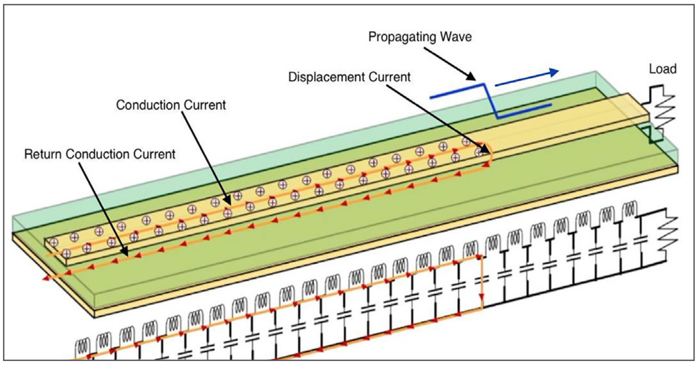

Figure 1 - A digital signal propagating along a microstrip with currents shown.

The signal's wave front travels at some fraction of the speed of light, as determined by the dielectric constant of the material, while the conduction current is comprised of a high density of free electrons moving at about 1 cm/second. The actual physical mechanism of near light speed propagation is due to a "kink" in the E-field, which propagates along the molecules of copper. Refer to *References 1, 2*, and *3* for further details.

The important thing is that this combination of conduction and displacement current must have an uninterrupted path back to the source. If it is interrupted in any way, the propagating electromagnetic wave will "leak" all around inside the PC board layers and cause "common mode" currents to form, which then couple to other signals (cross-coupling) or to "antenna-like structures", such as I/O cables or slots/apertures in shielded enclosures.

Most of us were taught the "circuit theory" point of view and it is important when we visualize how return currents want to flow back to the source. However, we also need to consider the fact that the energy of the signal is not only the current flow, but an electromagnetic wave front moving through the dielectric, or a "field theory" point of view. Keeping these two concepts in mind just reinforces the importance of designing transmission lines (signal trace with return path directly adjacent), rather than just simple circuit trace routing.

It is very important to note that all power distribution networks (PDNs) and high frequency signal traces are transmission lines and the energy is transferred as electromagnetic waves at about half the speed of light in normal FR4-type board dielectrics. We'll show what happens when the return path or return plane is interrupted by a gap in the next article. More on PDN design may be found in *Reference 4, 5,* and *6*.

Differential Mode versus Common Mode Currents

Referring to *Figure 2*, the differential mode current (in blue) is the digital signal itself (in this case, shown in a ribbon cable). As described above, the conduction current and associated return current flow simultaneously as the signal wave front moves along the transmission line formed by the microstrip and return plane.

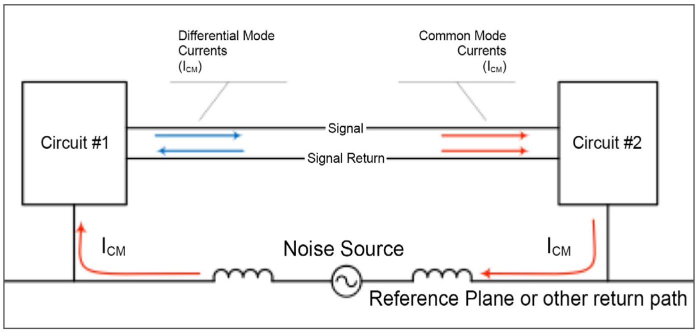

Figure 2 - An example of differential and common mode currents.

The common mode current (in red) is a little more complex in that it may be generated in a number of ways. In the figure, the impedance of the return plane results in small voltage drops due to multiple simultaneous switching noise (SSN) by the ICs. These voltage drops induce common noise currents to flow all over the return (or reference) plane and hence, couple into the various signal traces.

Besides SSN, common mode currents can also be created by gaps in return planes, poorly terminated cable shields, or unbalanced transmission line geometry. The problem is that these harmonic currents tend to escape out along the outside of shielded I/O or power cables and radiate. These currents can be very small, on the order of µA. It takes just 5 to 8 µA of current to fail the FCC class B test limit.

Summary

To summarize product design for EMI compliance, a properly designed PC board with adjacent return planes to all signals and PDNs, properly bonded I/O cable shields, well bonded shielded enclosures with minimal slots or gaps, and common mode filtering on all I/O and power cables for unshielded products is generally required for best EMI performance. Paying attention to these factors early in the design greatly reduces the risk of EMC and EMI compliance failures.

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For a number of years, Working Groups for EMC Society sponsored standards projects have met in parallel with the Technical symposium. This year, many standards related activities will take place as part of the Technical program. Proposals for standards related sessions are invited focusing on all aspects of standards contributions, including tutorial material, workshops on existing standards, novel contributions to standards projects or appraising the need for new standards.

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EMI TROUBLESHOOTING - STEP-BY-STEP

Kenneth Wyatt
Wyatt Technical Services LLC
ken@emc-seminars.com

In this article, I'll describe the steps I usually take to troubleshoot the top four EMI issues, conducted emissions, radiated emissions, radiated immunity, and electrostatic discharge. Of these, the last three are the most prevalent issues, with radiated emissions typically being the number one failure. If your product or system (EUT) has adequate power and I/O port filtering, conducted emissions and the other power line-related immunity tests are not usually an issue.

For your convenience, I've developed a list of recommended equipment useful for troubleshooting EMI. The download link is listed in Reference 1.

Originally published in the
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Conducted Emissions
This is usually not an issue given adequate power line filtering, however, many low-cost power supplies lack good filtering. Some "no name" brands have no filtering at all! The conducted emissions test is easy to run, so here you go.

Set up your spectrum analyzer as follows:

1. Frequency 150 kHz to 30 MHz
2. Resolution bandwidth = 10 or 9 kHz
3. Preamp = Off
4. Adjust the Reference Level so the highest harmonics are displayed and the vertical scale is reading in even 10 dB increments
5. Use average detection initially and CISPR detection on any peaks later
6. Internal attenuation - start with 20 to 30 dB at first and adjust for best display and no analyzer overload.
7. Set the vertical units to dBμV

I also like to set the horizontal scale from linear to log, so frequencies are easier to read out.

Obtain a Line Impedance Stabilization Network (LISN) and position it between the product or system under test and the spectrum analyzer. Note the sequence of connection below!

CAUTION: It's often important to power up the EUT prior to connecting the LISN to the analyzer. This is because large transients can occur at power-up and may potentially destroy the sensitive input stage of the analyzer. Note that the TekBox LISN has built-in transient protection. Not all do…you've been warned!

Power up the EUT and then connect the 50-Ohm output port of the LISN to the analyzer. Note the harmonics are usually very high at the lower frequencies and taper off towards 30 MHz. Be sure these higher harmonics don't overdrive the analyzer. Add additional internal attenuation, if required.

By comparing the average detected peaks with the appropriate CISPR limits, you'll be able to tell whether the EUT is passing or failing prior to formal compliance testing.

Ambient Transmitters
One problem you'll run into immediately is that when testing outside of a shielded room or semi-anechoic chamber, is the number of ambient signals from sources like FM and TV broadcast transmitters, cellular telephone, and two-way radio. This is especially an issue when using current probes or external antennas. I'll usually run a baseline plot on the analyzer using "Max Hold" mode to build up a composite ambient plot. Then, I'll activate additional traces for the actual measurements. For example, I often have three plots or traces on the screen; the ambient baseline, the "before" plot, and the "after" plot with some fix applied.

Often, its easier to narrow the frequency span on the spectrum analyzer down to zero in on a particular harmonic, thus eliminating most of the ambient signals. If the harmonic is narrow band continuous wave (CW), then reducing the resolution bandwidth (RBW) can also help separate the EUT harmonics from nearby ambients. Just be sure reducing the RBW doesn't also reduce the harmonic amplitude.

Another caution is that strong nearby transmitters can affect the amplitude accuracy of the measured signals, as well as create mixing products that appear to be harmonics, but are really combinations of the transmitter frequency and mixer circuit in the analyzer. You may need to use an external bandpass filter at the desired harmonic frequency to reduce the affect of the external transmitter. Although more expensive, an EMI receiver with tuned preselection would be more useful than a normal spectrum analyzer in high RF environments. Keysight Technologies and Rohde & Schwarz would be suppliers to consider. All these techniques are described in more detail in *Reference 3*.

Radiated Emissions
This is normally the highest risk test. Set up your spectrum analyzer as follows:

1. Frequency 10 to 500 MHz
2. Resolution bandwidth = 100 or 120 kHz
3. Preamp = On (or use an external 20 dB preamp if the analyzer lacks this)
4. Adjust the Reference Level so the highest harmonics are displayed and the vertical scale is reading in even 10 dB increments
5. Use positive peak detection
6. Set the internal attenuation = zero

Sometimes I prefer setting the vertical units from the default dBm to dBμV, so the displayed numbers are positive. This is also the same unit used in the test limits of the standards. I also like to set the horizontal scale from linear to log, so frequencies are easier to read out.

I perform my initial scan up to 500 MHz, because this is usually the worst case band for digital harmonics. You'll want to also record the emissions at least up to 1 GHz (or higher) in order to characterize any other dominant emissions. Generally speaking, resolving the lower frequency harmonics will also reduce the higher harmonics.

Near Field Probing
Most near field probe kits come with both E-field and H-field probes. Deciding on H-field or E-field probes depends on whether you'll be probing currents - that is, high di/dt - (circuit traces, cables, etc.) or high voltages - that is, dV/dt - (switching power supplies, etc.) respectively. Both are useful for locating leaky seams or gaps in shielded enclosures.

Start with the larger H-field probe (*Figure 1*) and sniff around the product enclosure, circuit board(s), and attached cables. The objective is to identify major noise sources and specific narrow band and broadband frequencies. Document the locations and dominant frequencies observed. As you zero in on sources, you may wish to switch to smaller-diameter H-field probes, which will offer greater resolution (but less sensitivity).

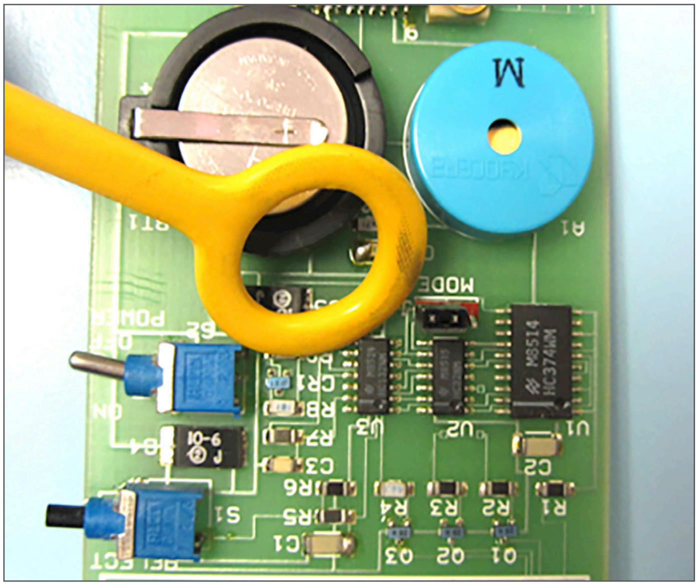

Figure 1. A near field probe is used to help identify potential sources of emissions.

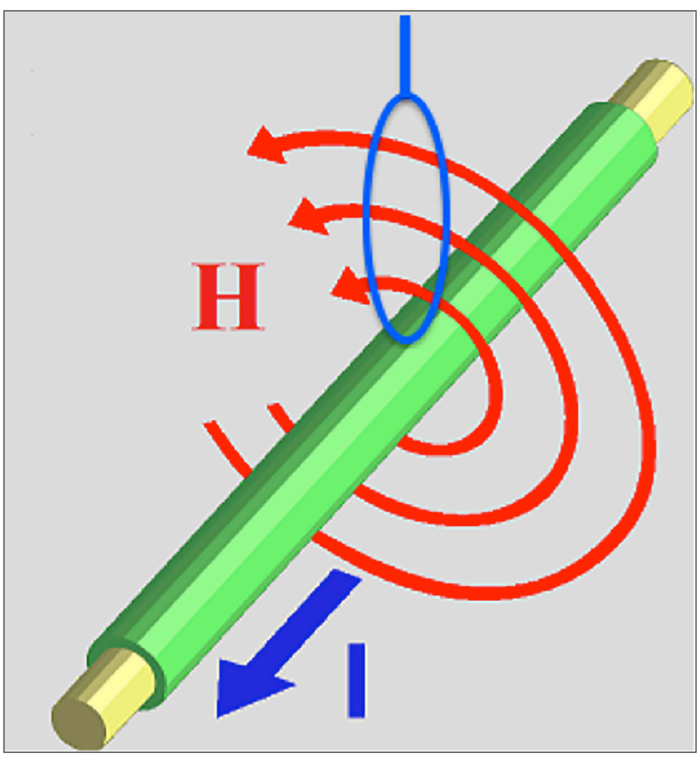

Figure 2. H-field probes offer the best sensitivity when oriented in relation to the circuit trace or cable, as shown. Figure, courtesy Patrick André.

Remember that not all sources of high frequency energy located on the board will actually radiate! Radiation requires some form of coupling to an "antenna-like" structure, such as an I/O cable, power cable, or seam in the shielded enclosure.

Compare the harmonic frequencies with known clock oscillators or other high frequency sources. It will help to use the Clock Oscillator Calculator, developed by my co-author, Patrick André. See the download link in *Reference 2*.

When applying potential fixes at the board level, be sure to tape down the near field probe to reduce the variation you'll experience in physical location of the probe tip. Remember, we're mainly interested in relative changes as we apply fixes.

Also, H-field probes are most sensitive (will couple the most magnetic flux) when their plane is oriented in parallel with the trace or cable. It's also best to position the probe at 90 degrees to the plane of the PC board. See *Figure 2*.

Current Probe

Next, measure the attached common mode cable currents (including power cables) with a high frequency current probe, such as the Fischer Custom Communications model F-33-1, or equivalent (*Figure 3*). Document the locations of the top several harmonics and compare with the list determined by near field probing. These will be the most likely to actually radiate and cause test failures, because they are flowing on antenna-like structures (cables). Use the manufacturer's supplied calibration chart of transfer impedance to calculate the actual current at a particular frequency. Note that it only takes 5 to 8 µA of high frequency current to fail the FCC or CISPR test limits.

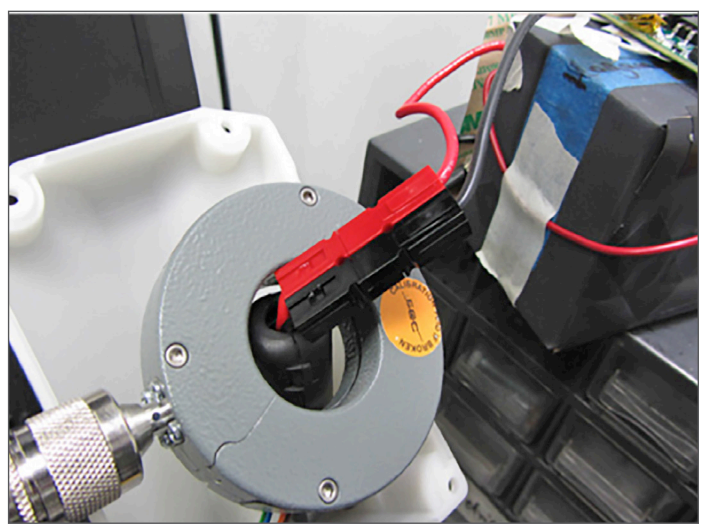

Figure 3. Use of a current probe to measure high frequency currents flowing on I/O and power cables.

It's a good idea to slide the current probe back and forth to maximize the harmonics. This is because some frequencies will resonate in different places, due to standing waves on the cable.

Its also possible to predict the radiated E-field (V/m) given the current flowing in a wire or cable, with the assumption the length is electrically short at the frequency of concern. This has been shown to be accurate for 1m long cables at up to 200 MHz. Refer to *Reference 3* for details.

Note on the Use of External Antennas
Note that there are two distinct goals when using external EMI antennas;

1. Relative troubleshooting , where you know areas of failing frequencies and need to reduce their amplitudes. A calibrated antenna is not required, as only relative changes are important. The important thing I that harmonic content from the EUT should be easily visible.
2. Pre-compliance testing, where you wish to duplicate the test setup as used by the compliance test lab. That is, setting up a calibrated antenna 3m or 10m away from the product or system under test and determining in advance whether you're passing or failing.

Pre-Compliance Testing for Radiated Emissions
If you're desiring to set up a pre-compliance test, (#2 above), then given a calibrated EMI antenna spaced 3m or 10m away from the EUT, you can calculate the E-field (dBµV/m) by recording the dBµV reading of the spectrum analyzer and factoring in the coax loss, external preamp gain (if used), any external attenuator (if used), and antenna factor (from the antenna calibration provided by the manufacturer). This calculation can then be compared directly with the 3m or 10m radiated emissions test limits using the formula:

E-field (dBµV/m) = SpecAnalyzer (dBµV) - PreampGain (dB) + CoaxLoss (dB) + AttenuatorLoss (dB) + AntFactor (dB)

For the purposes of this article, I'll focus mainly on the procedure for troubleshooting using a close-spaced antenna (#1 above) for general characterization of harmonic levels actually being radiated and testing potential fixes. For example, knowing you may be over the limit by 3 dB at some harmonic frequency means your goal should be to reduce that emission by 6 to 10 dB for adequate margin.

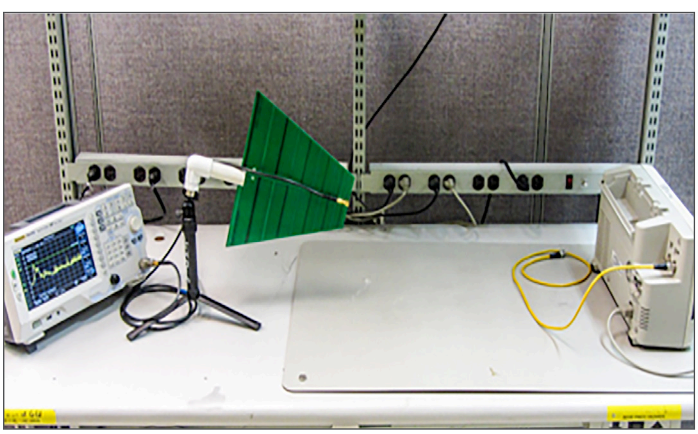

Figure 4. A typical test setup to measure actual radiated emissions while troubleshooting the causes.

Troubleshooting with a Close-Spaced Antenna
Once the product's harmonic profile is fully characterized, it's time to see which harmonics actually radiate. To do this, we use an antenna spaced at least 1m away from the product or system under test to measure the actual emissions (*Figure 4*). Typically, it will be leakage from attached I/O or power cables, as well as leakage in the shielded enclosure. Compare this data to that of the near field and current probes. Can you now determine the probable source(s) of the emissions noted?

Try to determine if cable radiation is the dominant issue by removing the cables one by one. You can also try installing a ferrite choke on one, or more, cables as a test. Use the near field probes to determine if leakage is also occurring from seams or openings in the shielded enclosure.

Once the emission sources are identified, you can use your knowledge of filtering, grounding, and shielding to mitigate the problem emissions. Try to determine the coupling path from inside the product to any outside cables. In some cases, the circuit board may need to be redesigned by optimizing the layer stack-up or by eliminating high speed traces crossing gaps in return planes, etc. By observing the results in real time with an antenna spaced some distance away, the mitigation phase should go quickly.

Common Issues
There are a number of product design areas that can cause radiated emissions:

1. Poor cable shield terminations is the top issue
2. Leaky product shielding
3. Internal cables coupling to seams or I/O areas
4. High speed traces crossing gaps in the return plane
5. Sub-optimal layer stack-up

Refer to the references for additional details on system and PC board design issues that can cause emissions failures.

Radiated Immunity
Most radiated immunity tests are performed from 80 to 1000 MHz (or, in some cases, as high as 2.7 GHz). Common test levels are 3 or 10 V/m. Military products can go as high as 50 to 200 V/m, depending on the operational environment. The commercial standard for most products is IEC 61000-4-3, whose test setup is quite involved. However, using some simple techniques, you can identify and resolve most issues quickly.

Handheld Radio
For radiated immunity, we generally start outside the EUT and use license-free handheld transmitters, such as the Family Radio Service (FRS) walkie-talkies (or equivalent) to determine areas of weakness. By holding these low power radios close to the product or system under test, you can often force a failure (*Figure 5*).

Hold the transmit button down and run the radio antenna all around the EUT. This should include all cables, seams, display ports, etc.

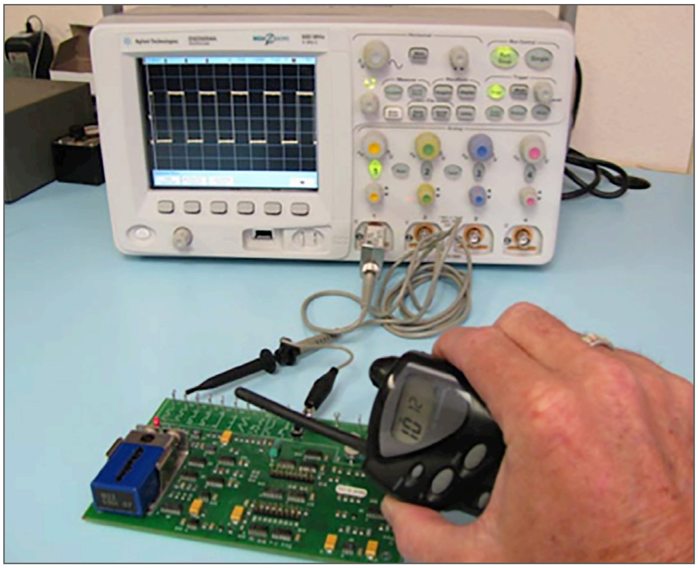

Figure 5. Using a license-free transmitter to force a failure.

RF Generator

It's very common that only certain frequency bands are susceptible and sometimes the fixed frequency handheld radios are not effective. In that case, I use an adjustable RF generator with attached large size H-field probe and probe all around at known failing frequencies. It also helps to probe the internal cables and PC board to determine areas of sensitivity. For smaller products, as in *Figure 6*, try using the smaller H-field probes for best physical resolution.

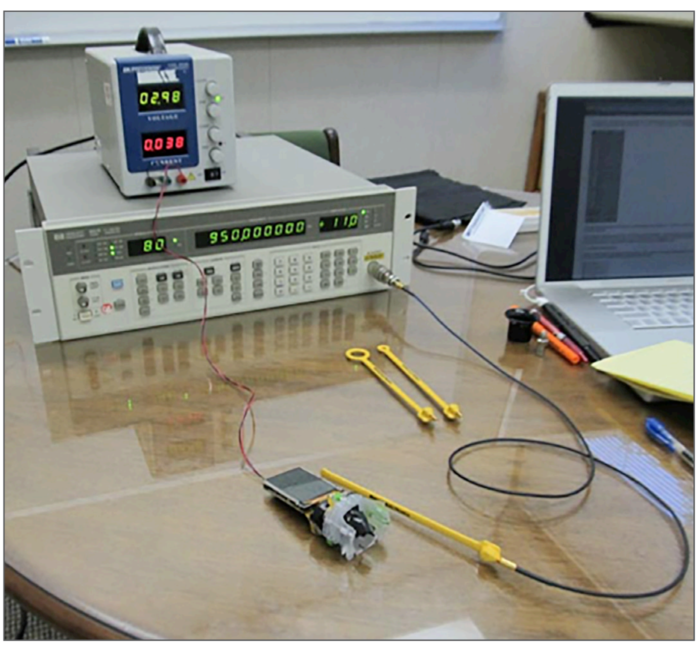

Figure 6. Using an RF generator and H-field probe to determine areas of sensitivity.

In place of the larger lab-quality RF generators, I also use a smaller USB-controlled RF synthesizer, such as the Windfreak SynthNV (or equivalent) with the near field probe. The SynthNV can produce up to +19 dBm RF power from 34 MHz to 4.4 GHz, so works well. This also fits into my EMI troubleshooting kit nicely. See Figure 7. You'll find a list of recommended generators in *Reference 1*.

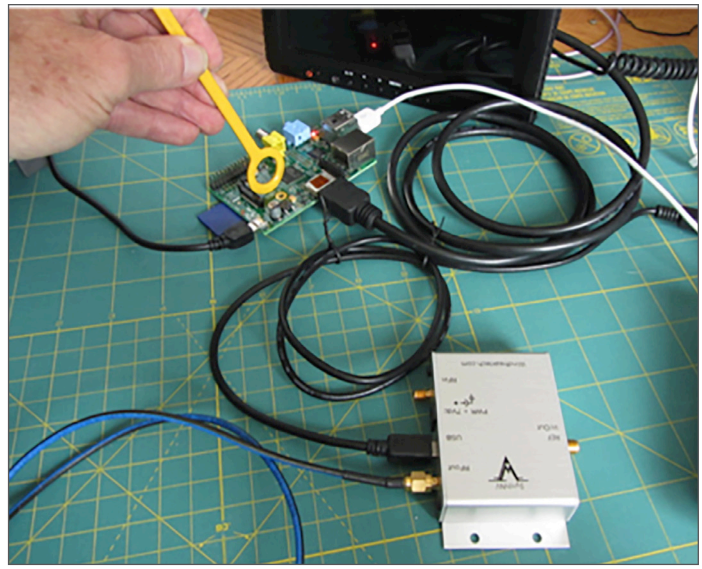

Figure 7. Using a small synthesized RF generator to produce intense RF fields around the probe tip.

Electrostatic Discharge

Electrostatic discharge testing is best performed using a test setup as described in the IEC 61000-4-2 standard. This requires a test table and ground planes of certain dimensions. The EUT is placed in the middle of the test table. I usually suggest replacing floor tiles with copper or aluminum 4 x 8-foot sheets, which will fit right into the spaces of the existing tiles (*Figure 8*).

Testing requires an ESD simulator, which is available from a number of sources. See *Reference 1*. I use the older KeyTek MiniZap, which is relatively small and can be adjusted to +/- 15 kV. There are several other suitable (and newer) designs.

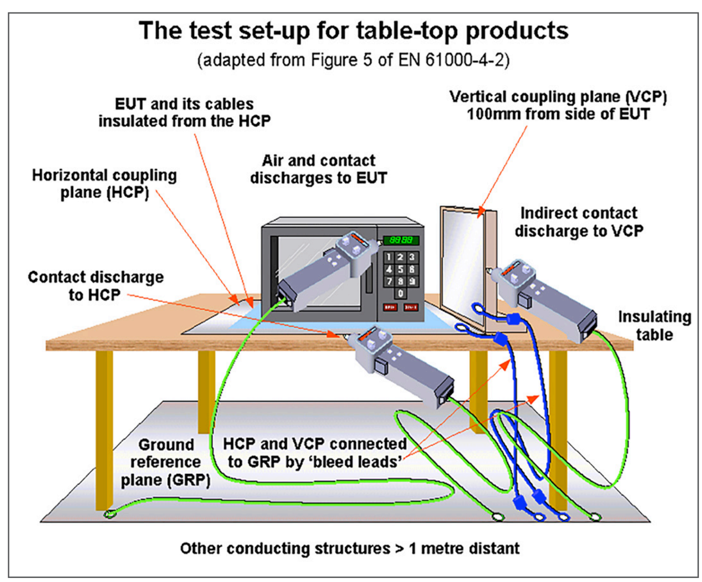

Figure 8. The ESD test setup according to IEC 6100-4-2. Image, courtesy Keith Armstrong.

ESD testing is rather complex as far as identifying the test points, but basically, there are two tests - air discharge and contact discharge. Use air discharge for all points where an operator could touch the outside of the EUT. Use contact discharge for all exposed metal where an operator could touch and discharge into. Test both positive and negative polarities. Most commercial tests require 4 kV contact discharge and 8 kV air discharge.

The test setup also includes horizontal and vertical coupling planes. Use the contact discharge tip into the coupling planes. These planes need a high-impedance discharge path to earth. See the IEC standard for details and exact test procedures.

Figure 9. A typical ESD simulator with air and contact discharge tips. It can produce up to +/- 15 kV.

Summary

By developing your own EMI troubleshooting and pre-compliance test lab, you'll save time and money by moving the troubleshooting process in-house, rather than scheduling time and the related cost and scheduling delays by depending on commercial test labs.

Most of the high-risk EMI tests are easily performed with low-cost equipment. The cost savings by performing troubleshooting at you own facility can mount up to hundreds of thousands of dollars and weeks or months of product delays.

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Author Bio

Kenneth Wyatt is president and principal consultant of Wyatt Technical Services LLC, as well as the senior technical editor for Interference Technology Magazine. He has worked in the field of EMC engineering for over 30 years with a specialty is EMI troubleshooting and pre-compliance testing. He is a co-author of the popular EMC Pocket Guide and RFI Radio Frequency Interference Pocket Guide. He also coauthored the book with Patrick André, EMI Trouble shooting Cookbook for Product Designers, with forward by Henry Ott. He is widely published and authored the EMC Blog hosted by EDN.com for nearly three years. Kenneth is a senior member of the IEEE and a long time member of the EMC Society.

He may be contacted at:
ken@emc-seminars.com
kwyatt@interferencetechnology.com

Check out his web site for more technical information, training schedules, and links:
http://www.emc-seminars.com

EMI TROUBLESHOOTING WITH REAL - TIME SPECTRUM ANALYZERS

Kenneth Wyatt
Wyatt Technical Services LLC
ken@emc-seminars.com

The latest tool for serious EMI troubleshooting or debugging has become the real-time (RT) spectrum analyzer. Because manufacturing costs have been decreasing, some RT analyzers are becoming more affordable than ever. In this article, I'll show you the advantages in using RT analysis for observing and troubleshooting unusual EMI.

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Download your copy at:
https://learn.interferencetechnology.com/2018-directory-design-guide/

Introduction

First, let's review the differences between the conventional swept and real-time spectrum analyzers.

Swept-Tuned Analyzer – The swept analyzer uses a tunable local oscillator (LO) in a standard superhetrodyne circuit. It can sweep over a specified frequency range and using a user-selected resolution (or "receiver") bandwidth. RF signals introduced to the input port are mixed with the local oscillator and the specified frequency span is display as RF power versus frequency. The only time data is captured is during the sweep time. After the frequency sweep, the captured data is processed and displayed. There is usually significant delay (or "dead" time) between sweeps, so its quite possible for the analyzer to miss capturing intermittent or fast-moving signals.

Real-Time Analyzer – A real-time analyzer uses a stationary LO, looks at narrow windows of bandwidth (real-time bandwidth), and digitizes the incoming spectrum. This digitized spectrum is stored in a time record buffer and held for processing by the FFT algorithm. Ideally, once digitized, FPGAs process FFTs at a rate equal, or faster, than the collection rate. However, this collection rate depends on the span and resolution bandwidth. The major difference between the swept-tuned analyzer and real-time analyzer is the sheer number-crunching ability of the real-time calculation, as well as a fast graphics processor, which allows for a data-dense display of various frequency-versus-time presentations and digital demodulation.

The advantages of a RT analyzer is the ability to capture RF pulses as short as 20 us, digital modulations, and other pulsing or fast changing signals. In addition, they can capture and process data much faster than swept analyzers – there's no need to wait seconds or minutes to capture a spectrum. This allows very fast troubleshooting, since you can see the result of fixes immediately.

Finally, the RT analyzers have an addition feature called a spectrogram (or "waterfall") display, where signals are shown versus time. This is a great feature allowing you to determine the timing of intermittent EMI.

I'll be using the Tektronix RSA306B (*Reference 1*) real-time USB-controlled spectrum analyzer with Tekbox Digital Solutions (*Reference 2*) near field probes for this article, but there are many other choices available.

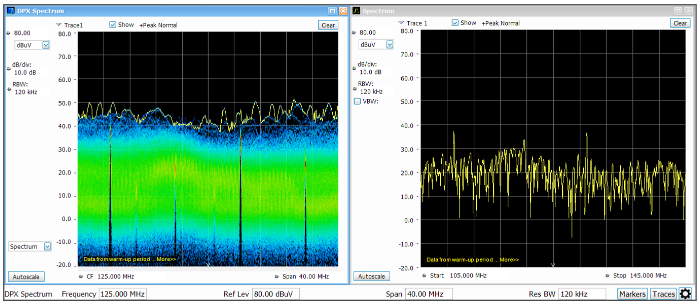

Figure 1 – An example where the broadband emissions from a motor controller completely mask a series of narrow band harmonics. You can see on the right that the standard swept analyzer has trouble capturing this broadband noise.

Figure 1 shows a typical advantage of the RT display over that of the swept display. Here, we see some broadband motor noise completely masking several narrow band harmonics. The swept analyzer has trouble capturing the motor noise, but we can see occasional captures indicating there was "something" there. Max Hold mode and waiting a while will help fill in the swept display, but then you'd miss seeing the narrow band emissions.

Most RT analyzers will also have optional EMI software that will help collect data or even perform pre-compliance testing for radiated and conducted emissions. For example, Tektronix offers their SignalVu-PC software with the RSA306B, but also recently announced their EMI troubleshooting and pre-compliance software for the RSA-series, called "EMCVu". EMCVu includes some impressive EMI troubleshooting and pre-compliance test features and can switch from one mode to the other quickly. It comes with pre-defined transducer factors (antenna and cable loss tables), CISPR and FCC limit lines, and easy report generation. In pre-compliance mode, it can scan the entire frequency range in a few seconds, numbering all the harmonics above the limit and within a certain margin to the limit. These captured harmonic signals can then be examined more closely and then switched over to troubleshooting mode to try various fixes.

Either SignalVu-PC or EMCVu will work fine for basic troubleshooting or debugging emission issues and I've actually used both for this article. If you also want pre-compliance test capability in-house (a wise decision) or more advanced troubleshooting tools, then you'll want to invest in EMCVu.

Three-Step Process for EMI Troubleshooting

I've developed a three-step process for EMI troubleshooting, which I'll briefly explain below. We'll use Tektronix' SignalVu-PC or EMCVu as an example, but several other companies sell similar compliance software. You'll want to download the free "2017 EMI Pre-Compliance Test Guide" from Interference Technology for more details on this troubleshooting process (*Reference 3*).

Step 1 – Use near field probes (either H- or E-field) to identify energy sources and characteristic emission profiles on the PC board and internal cables. Energy sources generally include clock oscillators, processors, RAM, D/A or A/D converters, DC-DC converters, and other sources, which produce fast-edged digital signals. If the product includes a shielded enclosure, probe for leaky seams of other apertures. Record the emission profile of each energy source.

Step 2 – Use a current probe to measure high frequency cable currents. Remember, cables are the most likely structure to radiate RF energy. Move the probe back and forth along the cable to maximize the highest currents. Record the emission profile of each cable.

Step 3 – Use a nearby antenna (I use a 1m test distance) to determine which of the harmonic content actually radiates. Catalog these harmonics and compare to the internal and cable measurements. This will help you determine the most likely energy sources that are coupling to cables or seams and radiating.

Analyze the Data

Remember that not all near field signals will couple to "antenna-like" structures and radiate. Use a harmonic analyzer tool (see *Reference 4*) to help identify harmonics belong to specific energy sources. Note that in many cases, two, or more, sources will generate the some (or all) the same harmonics. For example, a 25 MHz clock and 100 MHz clock can both produce harmonics of 100, 200, 300 MHz, etc. Oftentimes, you'll need to fix more than one source to eliminate a single harmonic. EMCVu includes some powerful data capture and documentation features that will help speed up the data collection process from steps 1 through 3.

After the harmonics are analyzed and you have identified the most likely sources, the next step is to determine the coupling path from source and out the product. Usually, it's the I/O or power cables that are the actual radiating structure. Sometimes, its leaky seams or apertures (display or keyboard, for example).

There are four possible coupling paths; conducted, radiated, capacitive, and inductive. The latter two (capacitive and inductive) are so-called; "near field" coupling and small changes in distance between source and victim should create large effects in radiated energy. For example, a ribbon cable routed too close to a power supply heat sink (capacitive coupling or dV/dt) and causing radiated emissions can be resolved merely by moving the ribbon able further away from the heat sink. The inductive coupling (di/dt) between a source and victim cable can also be reduced by rerouting. Both these internal coupling mechanisms (or similar PC board design issues) can lead to conducted (out power cables) or radiated (I/O or power cables acting as antennas, or enclosure seams/apertures) emissions.

In many cases, its simply poor cable shield bonding to shielded enclosures or lack of common-mode filtering at I/O or power ports that lead to radiated emissions.

How Can RT Analyzers Help Troubleshoot EMI?

So, let's turn our attention back to probing the PC board and cables. How often have you probed, troubleshot, and fixed a product only to have it fail at the compliance test facility? Many of today's products, especially mobile products, include on board DC-DC converters that produce a very broadband EMI spectrum out past 1 GHz that can impact the operation of cellular or GPS wireless receivers. In addition, digital processors can change emission characteristics with time or operating mode. Add wireless features and you have a myriad of potential energy sources that can change emission characteristics with time.

I'd like to demonstrate a some examples where swept analyzers might very well miss a bursting increase in emissions or fail to capture broadband EMI that is greater in amplitude than the usual narrow band harmonics we're all used to.

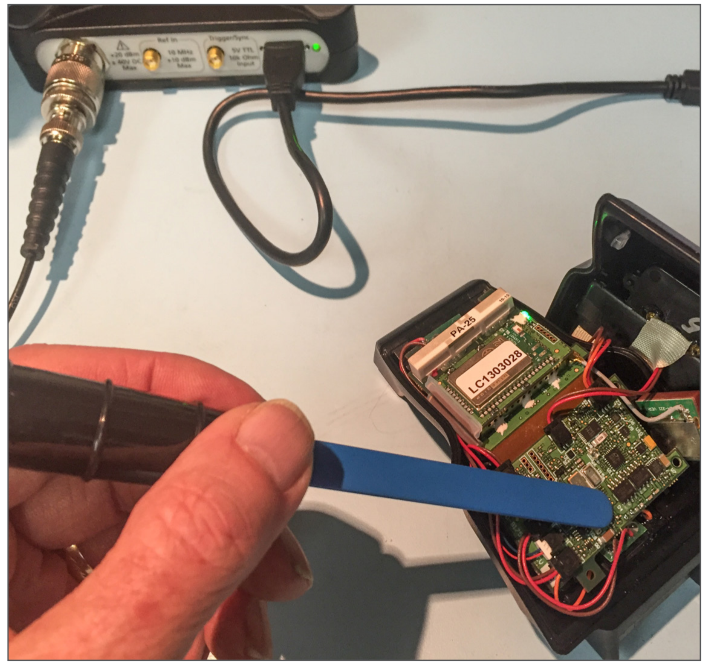

Figure 2 – Using a near field (H-field) probe on an on-board DC-DC converter in a small mobile device. I'm using the Tektronix RSA306B USB-controlled RT spectrum analyzer and Tekbox near field probe.

Example 1 – Pulsating Harmonic EMI

Most of the time, you'll find narrow band harmonics are relatively stable in amplitude. However, there are times when the amplitude can change, due to gated digital signals or different operating modes. If the harmonic peaks upward at the wrong time, it can lead to compliance failures.

Swept analyzers can easily miss these infrequent amplitude peaks. Placing the swept analyzer in "Max Hold" mode can help, but it could take several minutes to capture the peak of the emission. Even so, peaks can be missed, due to dead time in between scans.

RT analyzers, on the other hand are adept at capturing fast changing signals. Here's an example where I was measuring the narrow band low frequency emissions from an on-board DC-DC converter on a small mobile device (*Figure 2*).

In *Figure 3*, we're looking from 9 kHz to 10 MHz and we see the swept measurement is even having a hard tome capturing the regular peak emissions, while the RT measurement captures the peaks easily and even detects an occasional six dB pulsing increase in amplitude (as shown in the blue persistence display). That infrequent pulsing amplitude increase could easily cause a compliance failure should it couple out through conduction or radiation.

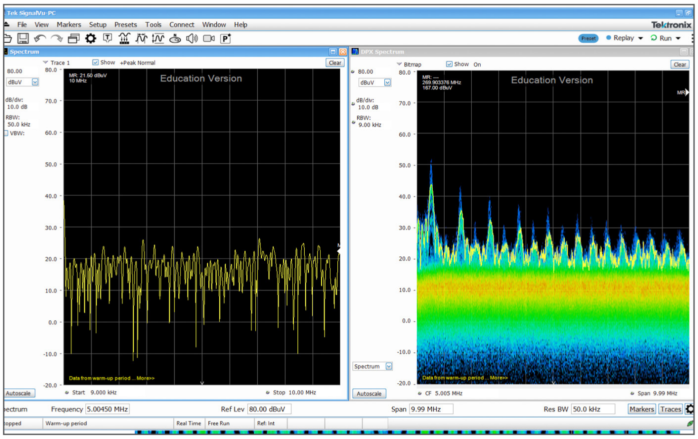

Figure 3 – Measuring the emissions from an on-board DC-DC converter and comparing swept (left) and real-time (right). Note the 6 dB peaks in the blue persistence display.

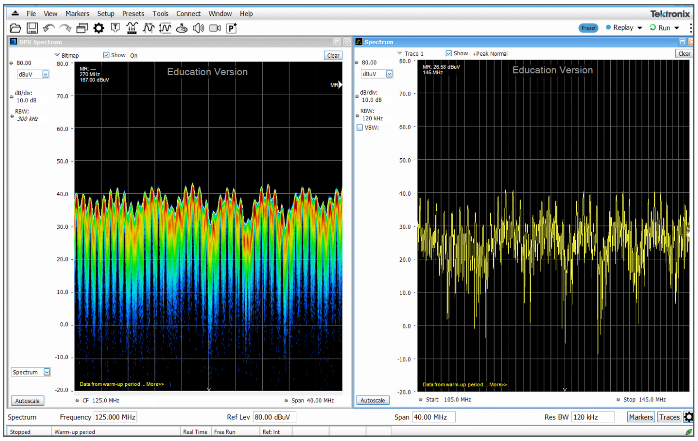

Figure 5 – Without moving the probe, we see "mode 2" from the DC-DC converter, which briefly appeared.

Example 2 – Identification of Emissions Due to Different Operating Modes

In this example, we're measuring that same DC-DC converter (*Figure 1*), but looking from 105 to 145 MHz, a frequent area of compliance failures due to radiated emissions. The surprising result was the three very different spectral responses, due to different operating modes of the mobile device. In some cases, the emission was about 25 dB higher than the swept measurement could capture. Now, would you be willing to take the risk that the swept measurement at the compliance test facility would either miss or manage to capture this should it couple out and radiate?

Figures 4, 5, and *6* show the three different spectral modes. Notice that the swept measurement managed to capture only two of the three spectrums. The near field probe was not moved during this sequence. Each mode was instantly viewable as the state changed from one mode to another.

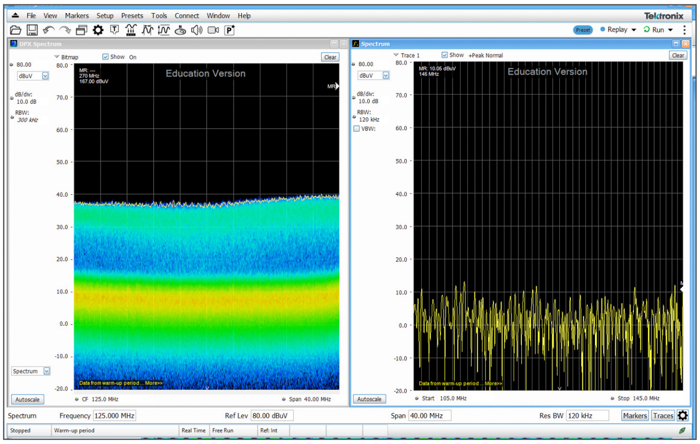

Figure 4 – Broadband emissions from the DC-DC converter looking from 105 to 145 MHz. The swept measurement on the right was unable to successfully capture this, except for an occasional burst. Max Hold mode would have helped, but would have taken at least a minute to "fill in" the display. But once the display was filled in, you may not have been able to see the following two very different modes in Figures 5 and 6.

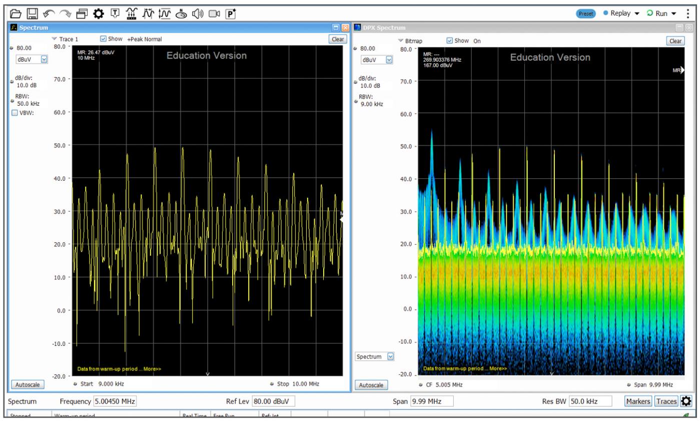

Figure 6 – Again, without moving the probe, we see "mode 3" with much increased narrow band emissions measuring about 10 dB higher than modes 1 and 2. This brief occurrence could have been the mode that would have resulted in a compliance failure, should the emission get coupled out and radiate.

Example 3 – Detection of Spurious Oscillation

In this example, we don't necessarily need the RT capture, but it does yield some interesting visual clues once we activate the spectrogram (waterfall) display feature.

The board being measured is a demo board from Picotest Technologies (*Figure 7*) and I discovered one of the op-amps produced an interesting bimodal series of spurious oscillations at about 150 MHz intervals. I was able to induce this oscillation by "switching out" the output capacitance.

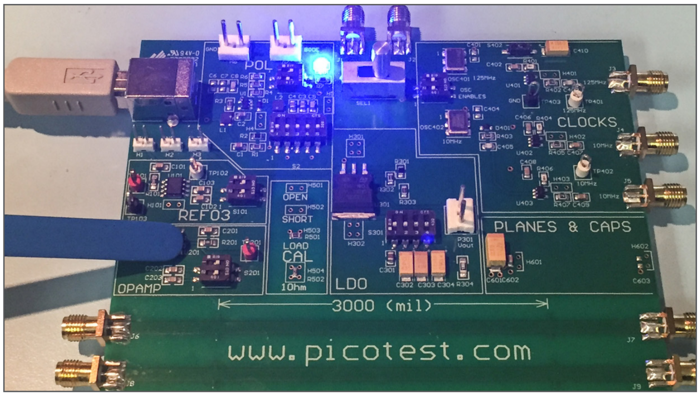

Figure 7 – Measuring an op-amp on the Picotest Technologies demo board.

It turns out that when the op-amp was unloaded capacitively, it produced a very interesting oscillation at near its open loop bandwidth (*Figure 8*). Examining the RT measurement on the right, we can see there's a distinct bimodal (two-frequency) display, along with some cool sideband emissions. The swept display on the left can only capture one of these two frequencies at a time, at best, as the oscillation is switching from one frequency to the other.

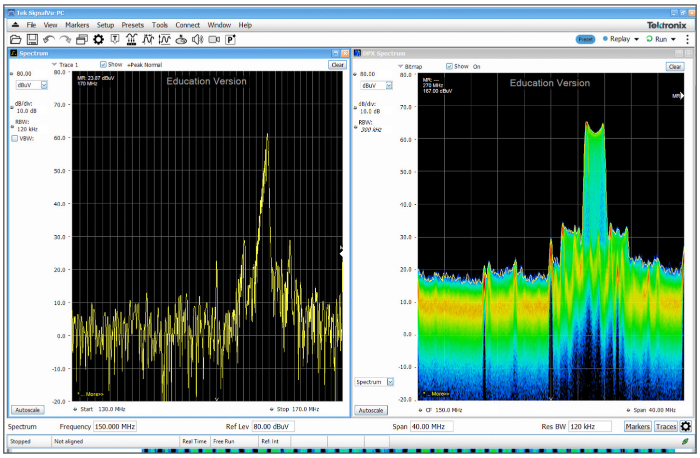

Figure 8 – Measurement of an interesting spurious oscillation of an op-amp. Note that the swept measurement on the left can only capture one of the bimodal states at a time, while the RT capture on the right is very detailed.

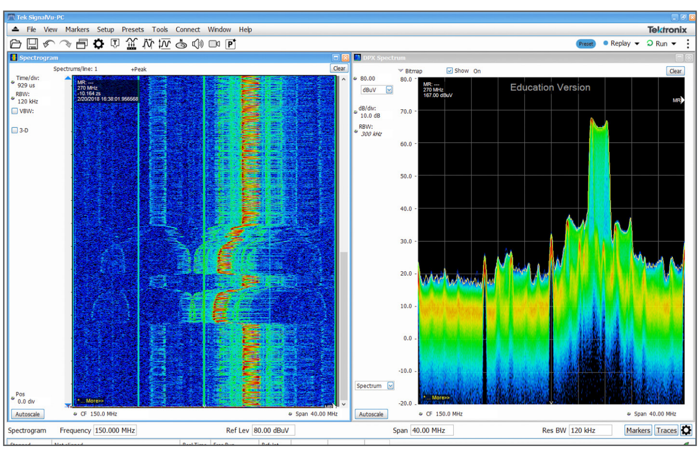

Figure 9 – Replacing the swept display with a spectrogram (frequency versus time), we can observe some interesting details (see text).

But let's analyze the "bi-modal-ness" a little closer by replacing the swept display with a spectrogram of frequency versus time.

One thing I noticed (and this is very common for spurious oscillations) is that placing my finger on the area of the op-amp changed the parasitic characteristics enough to shift the oscillation frequency quite a bit downward. You can see that shift in the spectrogram display in *Figure 9* as I touched my finger to the area twice.

The other thing to note is that you can now easily observe the switching between one oscillation frequency and the other in the "zig zag" pattern in the spectrogram. Note that the oscillation spends more time at the lower frequency, rather than the upper frequency. This is also indicated by the slightly higher amplitude of the left side of the double peak.

Conclusion

As technology continues to advance, we EMC engineers and product designers need to upgrade our usual analysis and pre-compliance test tools to stay one step ahead and be able to better capture and display the more unusual emissions expected. Real-time spectrum analyzers have already proven to be invaluable for EMI debug and troubleshooting. Advanced spectral analysis will be especially important as mobile devices continue to shrink and more products incorporate wireless and other advanced digital modes.

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Author Bio

Kenneth Wyatt is president and principal consultant of Wyatt Technical Services LLC, as well as the senior technical editor for Interference Technology Magazine. He has worked in the field of EMC engineering for over 30 years with a specialty is EMI troubleshooting and precompliance testing. He is a co-author of the popular EMC Pocket Guide and RFI Radio Frequency Interference Pocket Guide. He also coauthored the book with Patrick André, EMI Trouble shooting Cookbook for Product Designers, with forward by Henry Ott. He is widely published and authored the EMC Blog hosted by EDN.com for nearly three years. Kenneth is a senior member of the IEEE and a long time member of the EMC Society.

He may be contacted at:
ken@emc-seminars.com
kwyatt@interferencetechnology.com

Check out his web site for more technical information, training schedules, and links:
http://www.emc-seminars.com

EMC RADIATED EMISSION MEASUREMENTS AT 1/3/5/10/30 METERS

Daniel D. Hoolihan
danhoolihanemc@aol.com

Introduction

There are two principal types of emission measurements in the world of electromagnetic compatibility, conducted emission and radiated emission. The conducted emission measurements are either a voltage-capacitive tap type of measurement (typically on a power line) or they are a current-clamp type of measurement (typically on a signal line).

Radiated emission measurements are unique in that they must always state "the horizontal distance from the Equipment-Under-Test (EUT) to the receiving antenna" in order to compare the measured values to the appropriate regulatory limit. This horizontal distance, which is typically one, three, five, ten, or 30 meters, and the limits (both regulatory and standard-based) associated with those horizontal distances are the subject of this article.

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One-meter Measurements

There are two well-known EMC-measurement standards that reference a one-meter measurement distance from the EUT to the receiving antenna for radiated emissions. They are MIL-STD 461 and RTCA DO-160. There are other EMC standards that also use the one-meter horizontal-distance; one primary example is CISPR 25 – Electromagnetic Disturbances Related to Electric/Electronic Equipment on vehicles and Internal Combustion Engine Powered Devices.

First released in 1968, MIL-STD 461 has always specified a one-meter EUT-to-antenna distance; originally inside of a shielded room with bare walls and, then, in later revisions, inside of a shielded room with anechoic material on the walls. MIL-STD 461 (the latest version is MIL-STD-461G – December, 2015) is the standard used to test and qualify products sold to United States Military organizations and it has been widely duplicated in other countries' specifications for EMC of military electronic products.

RTCA DO-160 was first published in 1975 and is the EMC standard for commercial aircraft electronics in the United States and it is maintained by RTCA (an organization incorporated in Washington, DC). The latest version is RTCA-DO-160G, which was released in December of 2010. It's Section 21 addresses "Emission of Radio Frequency (RF) Energy" and it specifies a one-meter antenna distance inside of a shielded room with anechoic material (electromagnetic field absorbers) on the ceiling and about one-half of the wall surfaces. The European version of RTCA-DO-160 is EUROCAE ED-14 (EUROCAE is the European Organization for Civil Aviation Equipment; RTCA and EUROCAE work closely together and their standards are harmonized). CISPR 25 specifies a one-meter antenna distance to be used for radiated emissions from Components/Modules in an Absorber Lined Shielded Enclosure (ALSE).

The one-meter antenna distance has worked well for both Military Standard approved products and for Commercial Aviation approved products. A one-meter separation distance is a reasonable distance between an RF source and receptor of RF energy inside of a plane, a tank, or a ship. With the exception of CISPR 25 (where, again, a one-meter antenna distance is logical for closely located electronics in a vehicle), major measurement standards for terrestrial-based commercial products have conspicuously avoided a one-meter antenna measurement distance.

Three-Meter Measurements

Three-meter measurements are growing increasingly prevalent in the measurement world. They have been used by the United States Federal Communications Commission (FCC) for a number of years. Specifically, measurements of Class B digital devices (computers and similar devices) have been permitted at 3-meters since 1979 (FCC Docket 20780). The rationale for a three-meter measurement distance for Class B equipment was that small business computers (as a source of RF energy) would be closer to the potential receptor of energy (TV, radios, etc.) than a large Class A Computer. The simulated model of the source-receptor duality for Class B Computers was a business having a small computer and an apartment (3-meters away) having the TV or radio receiver.

Par. 15.109 (Radiated emission limits) of the FCC Rules says:

(a) **Except for Class A digital devices**, the field strength of radiated emissions from unintentional radiators at a distance of **3 meters** shall not exceed the following values:

Frequency of Emission (MHz)	Field Strength (microvolts/meter or dBuV/m)
30-88	100 uV/m or 40 dBuV/m
88-216	150 uV/m or 43.5 dBuV/m
216-960	200 uV/m or 46 dBuV/m
Above 960	500 uV/m or 54 dBuV/m

Three-meter measurements from 30-1000 MHz can be made in an Open Area Test Site or, more likely these days, in a three-meter semi-anechoic chamber due to the increasingly higher-ambient electromagnetic levels found in the environment. Fully anechoic rooms are also becoming more prevalent for 3-meter measurements.

It should be noted that three-meter measurements are also specified for radiated emission measurements for both Class A and Class B products above 1 GHz for both FCC Rules and International Standards for electromagnetic emissions.

Also, CISPR 32 – Edition 2.0 was published in 2015 and it allows Class A and Class B computers to be tested at a 3-meter horizontal distance from 30-1000 MHz.

Five-Meter Measurements

Radiated emission measurements made at a 5-meter horizontal antenna distance are being made in the commercial world. This is a "compromise" distance between 3-meters and 10-meters. The advantages to measurements made at 5-meters are that you can have a larger turntable in a 5-meter chamber and it is "easier" to meet the Volumetric Normalized Site Attenuation criteria for 3-meter distances in a larger 5-meter room.

However, at the present time, no standards specifically call out a 5-meter "standard" measurement distance. Limits specified at 3-meters or 10-meters are interpolated to a 5-meter distance using the inverse-distance fall-off guidance for a number of standards.

Ten-Meter Measurements

Many EMC technical experts consider the ten-meter measurement distance to be the "Gold Standard" in today's electromagnetic emission measurement world for Class A equipment. Ten-meter measurements are made at both Open Area Test Sites and in Semi-Anechoic Chambers. The semi-anechoic chambers are increasingly popular due to the steadily rising ambient levels in the real world because of digital TV and other new electronic developments.

Other advantages of the 10-meter antenna distance is that it allows a larger turntable to be used, and, therefore larger products can be tested with the receiving antenna in the "far-field" of the product's emanations. For example, at 10-meters, the EUT is one wavelength away from the antenna at 30 MHz, two wavelengths away at 60 MHz and three wavelengths away at 90 MHz. In contrast, equipment tested at 3-meters is not one wavelength away from the antenna until frequencies are at 100 MHz, two wavelengths away at 200 MHz, and three wavelengths away at 300 MHz.

Again, the FCC Rules are strongly stated in Par. 15.109 (Radiated emission limits) where it says:

(b) The field strength of radiated emissions from a Class A digital device, as determined at 10 meters, shall not exceed the following:

Frequency of Emission (MHz)	Field Strength (microvolts/meter or dBuV/m)
30-88	90 uV/m or 39 dBuV/m
88-216	150 uV/m or 42 dBuV/m
216-960	210 uV/m or 46.5 dBuV/m
Above 960	300 uV/m or 49.5 dBuV/m

NOTE - Several Asian countries require strict acceptance of 10-meter radiated emission measurements for Class A equipment when specified in their regulatory requirements based on international standards.

30-Meter Measurements

Thirty-meter measurements were the preferred measurement distance for Class A Digital Devices when the FCC rules were first released for 'computers' back in 1979.

The main reason for this was the CBEMA Report[1] released in 1977 in response to FCC Docket 20780[2]. The 1977 CBEMA report states "89 percent of receiving antennas found within 100 meters of commercial Electronic Data Processing/Office Equipment installations can be expected to be 30 meters or more from the installations." **Therefore, the CBEMA report chose "30 meters" as a reasonable control distance for radiated emission limits from Class A computers.**

Also, the FCC imposed rules at 30 meters (approximately 100 feet). In a historical article[3] by Herman Garlan, Chief of the Radio Frequency (RF) Devices Branch in 1973, he states, "The rules then in effect (for operation with a duty cycle) permitted a field-strength level of 50 uV/m at 100 feet (30 meters) on frequencies between 88-108 MHz."

Also, in the 1970s, the German VDE testing authorities used a 30-meter test distance for much of their testing[4].

Problems with relatively high-ambient levels from 30 MHz to 1000 MHz at 30 meters made it very difficult to make measurements. In addition, the antenna mast had to be 6 meters high at 30 meters, which was a challenge for EMC test labs. Normalized Site Attenuation (NSA) was also a technical challenge for 30-meter test sites; it was achievable but time-consuming and more complex than NSA at 10 meters or 3 meters.

Because of the above difficulties, in the early 1980s the FCC released Docket 80-284, which eventually changed the preferred test distance for Class A digital devices to 10 meters. **So, in the United States, the 10-meter distance for Class A devices has been the dominant distance for the last 35 years.**

NOTE - There are strong technical arguments for using a 30-meter test distance for frequencies BELOW 30 MHz due to the longer wavelengths of the electromagnetic energy at lower frequencies.

International Standards

Despite a number of changes to FCC Rules since the first publication of this article in 2010, Part 15 of the FCC rules still states in 15.109 (g):

"As an alternative to the radiated emission limits shown in paragraphs (a) and (b) of this section, digital devices may be shown to comply with the standards contained in the **Third Edition** of the International Special Committee on Radio Interference (CISPR), Pub. 22, "Information Technology Equipment (ITE) – Radio Disturbance Characteristics – Limits and Methods of Measurement."

The **Third Edition of CISPR 22 (1997)** has the following limits:

Table 5 – Limits for Radiated Disturbance of Class A ITE at a measuring distance of 10 meters

Frequency Range – MHz	Quasi-Peak Limits – dBuV/m
30-230	40 (=100 uV/m)
230-1000	47 (=224 uV/m)

NOTE – The CISPR 32 – 2015 limits for Class A and Class B are same as CISPR 22 – 1997 at 10 meters.

Table 6 – Limits for Radiated Disturbance of Class B ITE at a measuring distance of 10 meters

Frequency Range – MHz	Quasi-Peak Limits – dBuV/m
30-230	30 (=32 uV/m)
230-1000	37 (=71 uV/m)

If we compare the FCC and CISPR 22 (1997)/CISPR 32 (2015) limits at <u>10 meters</u> for Class A equipment, we have the following table:

Frequency of Emission (MHz)	FCC – CLASS A (microvolts/meter)/(dBuV/m)	CISPR 22 – CLASS A CISPR 32 – CLASS A (microvolts per meter)/(dBuV/m)
30-88	90/39	100/40
88-216	150/43.5	100/40
216-230	210/46.5	100/40
230-960	210-46.5	224/47
Above 960	300/49.5	224/47

If we compare the FCC and CISPR 32 (2015) limits at 3 me-

ters for Class B equipment, we have the following table:

Frequency of Emission	FCC – CLASS B (microvolts/meter)/(dBuV/m)	CISPR 32 – CLASS B (microvolts per meter)/(dBuV/m)
30-88	100/40	100/40
88-216	150/43.5	100/40
216-230	200/46	100/40
230-960	200/46	224/47
Above 960	500/54	224/47

The two sets of limits (FCC and CISPR 22/CISPR 32) are reasonably close for Class A equipment at 10 meters and further apart for Class B equipment at 3-meters.

Inverse Distance Fall-Off

The inverse distance fall-off theory, also called the 1/r (1/d) theory, assumes a small source in a free-space (free-field) environment. **In general, these two conditions (small source and free-space) are not met in a typical EMC measurement.**

Most products have lengths and widths so they are not necessarily a "small source", for example, a table-top product is placed on a non-conductive table 0.8 meter above the ground plane and the power cord from EUT starts at the ground plane and reaches up to the EUT. The non-conductive table has a nominal size of 1.0 meter wide and 1.5 meters long. The product under test is usually smaller than the table but it is possible for it to be bigger than the standard table.

The ground plane is typically a solid metal floor or a metallic screen with small openings. In both cases, a reflected wave from the ground plane complicates the measurement of the radiated fields from the EUT. There have been a number of technical studies on the fall-off of electromagnetic fields from measurements close to a product versus a regulatory limit at a further distance from the product. We will look at a number of those studies in this paper.

Technical Studies Justifying Inverse Distance Fall For Real Products

Note: The author was unable to find any technical paper that justifies an inverse-distance fall-off for real products in an Open Area Test Site or a Semi-Anechoic Chamber especially for distances below 10 meters and frequencies between 30 and 1000 MHz.

Technical Studies Questioning Inverse Distance Fall For Real Products

One of the first papers on "Falloffs" was written by William E. Cory and Frank C. Milstead in 1969[6]. It stated: "Propagation predictions in the near field, while less accurate, can be made to within about 10 db."

Albert A. Smith, Jr. wrote a paper in 1969[7], which modeled surface waves and space waves and found a complex relationship **below 100 MHz**. However, the paper goes on to say "Above approximately 100 MHz the space wave predominates for 'source and receiving heights of 1 meter' and the induction fields are negligible for 'antenna to EUT distances' greater than one meter."

Herman Garlan's paper[8] says in the "History of Part 15" section "The original low-power rule, the λ/2π rule, was adopted in 1938. This rule provided a reasonable operating standard on frequencies up through the AM broadcast band – up to 1600 kHz. This standard was still usable up to about 10 MHz where the λ/2π rule permits a field of 15 uV/m at about 5 meters or 16 feet. **While this standard served the needs of 1938, by the end of World War II, in 1945, it was hopelessly inadequate."**

The CBEMA paper was published in 1977; it was a comprehensive review of the interference potential of large computers. It says, "A practical site that allows measurements at the minimum test distance of 3 meters is shown in Figure 10-3. **Results of measurements in such practical test sites at varying distances between the equipment being tested and the measurement antenna, have been found to be within +/- 6 dB of those predicted using a 20 dB/decade fall-off relationship between the equipment and the antenna."**

Yet another paper was published in 1980 by Robert F. German and Ralph Calcavecchio[8]. This paper says "It is generally accepted that EMI radiated from large equipment should be measured at a distance of 30 meters. Measurements in the 30-1000 MHz frequency range at this distance usually are in the far-field of the source. However, 'due to ambient conditions' it is desirable to allow measurements to be made at distances of 3, 10, or 30 meters. It will be seen that, when appropriate assumptions are made, a measurement technique can be identified that relates measurements made at different distances by the 1/r attenuation factor of free space propagation." **The paper goes on to say "An EMI source is simulated by an electrically short dipole antenna. Actual EMI sources may be more complex and the topic of future work." Thus, the paper concludes 1/r works for "electrically short dipoles."**

Another paper from two engineers who worked at IBM[9], concluded **"The radiation from more than 25 different products showed a great variation from the 20 dB attenuation often assumed between three and 30 meter field strength levels."** It stated further "These products varied in maximum linear dimension from one to 10 meters." Also, the paper had three E-field falloff figures; "In all three falloff figures, it is noted that the radiated field at few frequencies attenuate at a rate of 20 dB per decade distance). This does not contradict the theoretical 20 dB falloff in free space between two points in the far-field located at a distance ratio of 10 to 1 away from a point source or from a dipole antenna small relative to the wavelength radiated. **In fact, a very large source (see Figure 8b) could in the extreme show a falloff approaching 0-dB because it contains a large number of geometrically distributed sources, both horizontally and vertically.** The fields from such multiple sources superimpose and may generate an almost plane wavefront (a plane wavefront exhibiting 0-dB falloff)."

Another paper[10], by Arlon T. Adams, Yehuda Leviatan, and Knut S. Nordby, covered a study concerned with the near fields of computer products. The study states that "The measurement distances of 3 to 30 meters may lie in the near or the far field, depending on the dimensions of the product and the frequencies emitted." Furthermore, the study says, **"In other words, the average slope in the oscillatory region is less than 20 dB per decade (it is about 10 dB per decade.) In other words, a product just meeting FCC rules at 3-m distance may exceed the rule when measured at 30 m. Thus, measurements made at short distances and then normalized to larger distances will yield far-fields smaller that they should be."** An additional paper by Adams and Nordby[11] reemphasized the above points.

In 1987, there was an article published in the 1987 IEEE International Symposium on EMC record[12] by J. D. Gavenda concerning vertical dipole sources in EUTs. **His paper stressed the point that vertical electrical fields are also produced off the end of a horizontal electric dipole, and broadside to a horizontal magnetic dipole.** The paper states "In free space at distances large compared with the wave-length and with the maximum dimensions of the EUT, the field strength falls off inversely with distance. However, the presence of a conducting ground plane causes reflected signals, which interfere constructively or destructively, depending on height above the ground plane and frequency, with the direct signal. **This invalidates any simple inverse-distance falloff rule, so correction factors must be used in the extrapolations."** In the paper, he has a falloff figure for a vertical dipole FROM 3 TO 10 meters that is a shallow-v-shaped with a only a 7 dB falloff from 30 to 100 MHz, a mere 4 dB falloff from 100 to 300 MHz, and, then, back to about a 7 dB falloff from 300 – 1000 MHz.

A very well known and respected paper was written in 1987 by Joseph DeMarinis of Digital Equipment Corporation[13]. One of the goals of this paper was the "Prediction and Measurement of correlation errors between 3-meter and 10-meter site distances and development of bands of confidence around such correlation." In its Introduction, the paper says **"It is well known that signal falloff versus site distance does not follow the 1/distance rule which is proscribed by the regulatory standards and that very large correlation errors can exist between test results taken at different distances. It was of particular interest to the project at hand, to try to understand the relationship between 3-meter and 10-meter sites."** The resulting data of the study showed a falloff of only 4 to 9 db from 30-200 MHz for vertical signals and a falloff between 9 and 14 dB for horizontal signals. From 200- 1000 MHz, the falloff for vertical signals ranged from 3 to 11 db and for horizontal signals it ranged from 8 to 13 db. All of this data, predicted and actual, was for Open Area Test Sites.

In 1993, three engineers from Austria wrote a paper on radiated emission testing at 3 meters[14]. This paper investigated a difference in extrapolation factors (0 db/decade in CISPR 11 and 20 dB/decade in CISPR 22) that existed at that time. Measurements were made at an Open Area Test Site and showed a range of falloff from 1 to 18 dB from a setup representing a typical personal computer. The paper presented worst-case extrapolation factors, for 3 and 10-meter test results, for both horizontal and vertical polarizations.

Another paper in 1996 by Christopher I. Holloway and Edward F. Kuester[15] looked at the comparison of OATS and semi-anechoic chambers. It stated that by looking at site attenuations of the two venues an equivalent comparison could be made. It concluded that **"This comparison is generally quite good at frequencies higher than 300 MHz, but at lower frequencies (30 -300 MHz), large discrepancies are often observed due to reflections from the chamber walls."**

Finally, a paper given in 2009 by Blankenship, Arnett, and Chen described another perspective on looking at the falloffs from 3 to 10 meters[16]. This paper also predicted a complicated falloff curve for signals between 3 and 10 meters and it was based on testing in semi-anechoic chambers.

Conclusions and Recommendations

It can be seen that over the past forty years that the measurement of radiated emissions from electronic equipment has been an active topic.

The military and commercial avionics, as well as automotive, products have consistently used (and continue to use) a one-meter antenna distance for radiated emission. However, they have made improvements in the shielded-room locale by adding anechoic material to the ceiling and, at least, part of the wall surfaces in the chamber thus reducing reflections and increasing the accuracy of the test results over the past five decades.

It is also observed that there has been a trend over the last forty years towards making measurements on commercial products at antenna distances closer and closer to the Equipment Under Test. We have gone from an environment of making measurements at 30 meters on Class A commercial electronic products to an environment of making measurements at 3 and 10 meters.

The risk with moving closer to the product under test is that the receiving antenna can be immersed in the near-field environment of the EUT. When this happens, and it does at various distances and frequencies depending on the size and internal sources in the product, **predicting falloffs of electromagnetic energy with the inverse distance falloff formula (1/r distance factor) does not work and the fields measured at distances further from the product will, in general, be at a higher amplitude than that predicted with a 1/r falloff.**

Class B Products Tested at 3-meters

Since Class B products are already commonly tested at 3 meters for FCC regulations from 30 MHz to 1000 MHz and Class A and B products are tested at frequencies above 1000 MHz at 3-meters both in the USA and worldwide, it is obvious that 3-meter measurements are widely accepted around the world.

If Class B products are tested at 3 meters as per CISPR 32, there would be no need for discussions relative to falloffs from 3 to 10 meters and the USA and International limits are very close which may lead to the desired goal of "harmonization."

Class B Products Tested at 5-meters
Class B products tested at 5 meters need to be investigated further as to their falloffs since there has been a limited amount of research done on the falloffs of fields from 3 to 5 meters and 5 to 10 meters over the frequency range 30 -1000 MHz.

Class A Products Tested at 10-meters
One alternative to the Class A issue is to mandate all Class A products be tested at 10 meters with no exceptions. Then, there would be no falloff debates since Class A products could not be tested at a closer distance.

However, if industry would like to test Class A products at 3-meters, as per the latest version of CISPR 32, there should be a correction factor applied to handle that situation. It is probably not 0 db (as was used in CISPR 11 in 1998) and it is probably not 10 dB (as used in CISPR 32 in 2015). It is some factor between those two theories and it should be frequency dependent.

A proposal along those lines would be a correction factor (not equal to the widely accepted 10 dB) that would be added to the 10-meter regulatory limit when the product is tested at three meters. As a first estimate, the following correction factors (instead of a de facto +10 dB) are proposed:

30 – 100 MHz - + 6 db

100 -300 MHz - + 3 dB

300 – 600 MHz - + 6 db

600 – 900 MHz - + 7 dB

900 – 1000 MHz - + 8 db

So, for example, at 120 MHz, the limit would be 40 plus 3 or 43 dBuV/m (instead of 40 plus 10 or 50 dBuV/m) when a Class A EUT is measured at a 3-meter antenna distance. (See Table A.2 of CISPR 32).

These proposed correction numbers are consistent with references[12] and[16]. This set of correction factors would cover the vertical field falloffs and would be even more conservative for the horizontal field falloffs (which are closer to the 1/r falloff curve).

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DESIGN FOR EMC COMPLIANCE ESSENTIALS

Kenneth Wyatt
Wyatt Technical Services
ken@emc-seminars.com

Introduction

While unrealistic to discuss all aspects of product design in a single article, I'll try to describe the most common design issues I find in the hundreds of client products I've had a chance to work on. These issues generally include PC board design, cables, shielding, and filtering. More detailed information may be found in the Reference section below.

As previously mentioned, the top three product failures I run into include (1) radiated emissions, (2) radiated susceptibility, and (3) electrostatic discharge. Other failures can include things like conducted emissions, electrically fast transient, conducted susceptibility, and electrical surge. Most of these last items are also the result of the same poor product designs, which cause the top three failures.

NOTE: I prefer to avoid the word "ground" in this article or in my consulting practice. The reason is that there are too many misinterpretations, which can also lead to EMC failures. It's much more clear to use power and power return, and signal and signal return - or just "return plane" or reference plane. Finally, cable shields or shielded enclosures are "bonded" together - not "grounded". The only exception is the so called "safety ground" or earth ground. But these have nothing at all to do with proper EMC design - just personal safety against electrical shock. I suppose the one exception would be the earth ground connection on a three-wire power line filter. Also, occasionally, there will be an earth ground on a PC board - especially for power supplies, but again, connecting a product or system to earth ground will not improve EMI, due to the very high inductance (length) of the wire.

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PC Board Design

The single most important factor in achieving EMC/EMI compliance revolves around the printed circuit board design. It's important to note that not all information sources (books, magazine articles, or manufacturer's application notes) are correct when it comes to designing PC boards for EMC compliance - especially sources older than 10 years, or so. In addition, many "rules of thumb" are based on specific designs, which may not apply to future or leveraged designs. Some rules of thumb were just plain lucky to have worked.

PC boards must be designed from a physics point of view and the most important consideration is that high frequency signals, clocks, and power distribution networks (PDNs) must be designed as transmission lines. This means that the signal or energy transferred is propagated as an electromagnetic wave. PDNs are a special case, as they must carry both DC current and be able to supply energy for switching transients with minimal simultaneous switching noise (SSN). The characteristic impedance of PDNs is designed with very low impedance (0.1 to 1.0 Ohms, typically). Signal traces, on the other hand, are usually designed with a characteristic impedance of 50 to 100 Ohms.

The previous article introduced the concept of the circuit theory and field theory viewpoints. A successful PC board design accounts for both viewpoints. Circuit theory suggests that current flows in loops from source to load and back to the source. In many cases of product failure, the return path has not been well defined and in some cases, the path is broken. Breaks or gaps in the return path are major causes of radiated emissions, radiated susceptibility, and ESD failures.

Correspondingly, electric fields on PC boards exist between two pieces of metal, such as a microstrip over a return plane (or trace). If the return path is broken, the electric field will "latch on" to the next closest metal and will not likely be the return path you want. When the return path is undefined, then the electromagnetic field will "leak" throughout the dielectric and cause common mode currents to flow all over the board, as well as cause cross-coupling of clocks or other high speed signals to dozens of other circuit traces within that same dielectric.

Figure 1 shows a propagating wave within the dielectric between the signal trace and return plane (or trace). This shows both the conduction current flowing in the signal trace and back on the return plane (or trace) and the displacement current "through" the dielectric. The signal wave front travels at some fraction of the speed of light as determined by the dielectric constant. In air, signals travel at about 12 inches per nanosecond. In the typical FR4 dielectric, the speed is about half that at 6 inches per nanosecond. Refer to *Reference 1, 2,* and *3* for more information on the physics of signal propagation through PC boards.

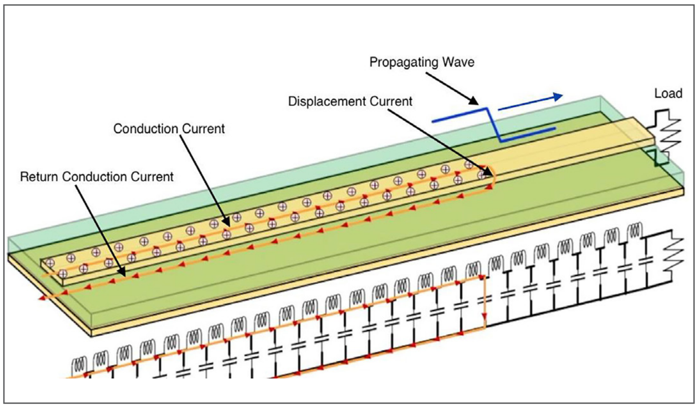

Figure 1 - A propagating wave along a microstrip with reference plane. Figure, courtesy Eric Bogatin.

In order to satisfy both the circuit and field theory viewpoints, we now see the importance of adjacent power and power return planes, as well as adjacent signal and signal return planes. PDN design also requires both bulk and decoupling "energy storage" capacitors. The bulk capacitors 4.7 to 10 µF, typ.) are usually placed near the power input connector and the decoupling capacitors (1 to 10 nF, typ) nearest the noisiest switching devices - and most importantly, with minimal trace length connecting these from the power pins to signal return plane. Ideally, all decoupling capacitors should be mounted right over (or close to) the connecting vias and multiple vias should be used for each capacitor to reduce series inductance.

Signal or power routed referenced to a single plane will always have a defined return path back to the source. *Figure 2* shows how the electromagnetic field stays within the dielectric on both sides of the return plane. The dielectric is not shown for clarity.

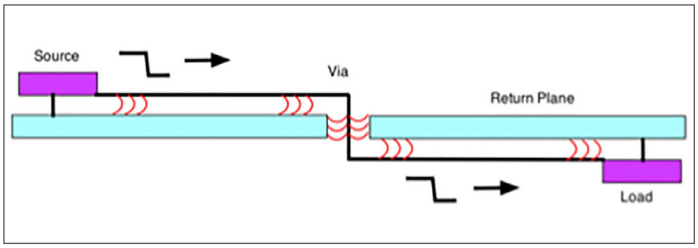

Figure 2 - A signal trace passing through a single reference plane.

On the other hand, referring to *Figure 3*, if a signal passes through two reference planes, things get a lot trickier. If the two planes are the same potential (for example, both are return planes), then simple connecting vias may be added adjacent to the signal via. These will form a nice defined return path back to the source.

If the two planes are differing potentials (for example, power and return), then stitching capacitors must be placed adjacent to the signal via. Lack of a defined return path will cause the electromagnetic wave to propagate throughout the dielectric, causing cross coupling to other signal vias and leakage and radiation out the board edges as shown.

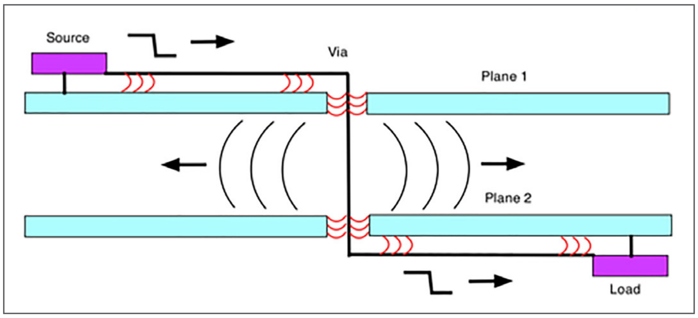

Figure 3 - A signal trace passing through two reference planes. If the reference planes are the same potential (signal or power returns, for example), then stitching vias next to the signal via should be sufficient. However, if the planes are different potentials (power and return, for example), then stitching capacitors must be installed very close to the signal via. Lack of a defined return path will cause the electromagnetic field to leak around the dielectric, as shown, and couple into other signal vias or radiate out board edges.

For example, let's take a look at a poor (but very typical) board stack-up that I see often. See *Figure 4*.

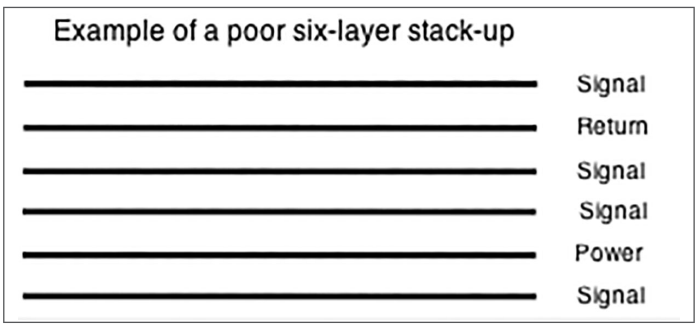

Figure 4 - A six-layer board stack-up with very poor EMI performance.

Notice the power and power return planes are three layers apart. Any PDN transients will tend to cross couple to the two signal layers in between. Similarly, few of the signal layers have an adjacent return plane, therefore, the propagating wave return path will jump all over to whatever is the closest metal on the way back to the source. Again, this will tend to couple clock noise throughout the board.

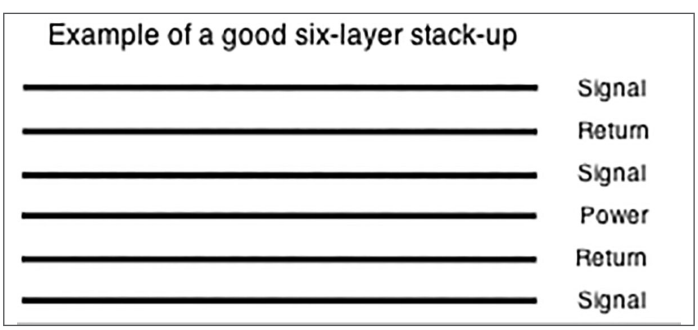

Figure 5 - A six-layer board stack-up with good EMI performance. Each signal layer has an adjacent return plane and the power and power return planes are adjacent.

A better design is shown in *Figure 5*. Here, we lose one signal layer, but we see the power and power return planes are adjacent, while each signal layer has an adjacent signal (or power) return plane. It's also a good idea to run multiple connecting vias between the two return planes in order to guarantee the lowest impedance path back to the source. The EMI performance will be significantly improved using this, or similar designs. In many cases, simply rearranging the stack-up is enough to pass emissions.

Note that when running signals between the top and bottom layers, you'll need to include "stitching" vias between the return planes and stitching capacitors between the power and power return planes right at the point of signal penetration in order to minimize the return path. Ideally, these stitching vias should be located within 1 to 2 mm of each signal via.

Other Tips - Other design tips include placement of all power and I/O connectors along one edge of the board. This tends to reduce the high frequency voltage drop between connectors, thus minimizing cable radiation. Also, segregation of digital, analog, and RF circuits is a good idea, because this minimizes cross coupling between noisy and sensitive circuitry.

Of course, high-speed clocks, or similar high-speed signals, should be run in as short and as direct a path as possible. These fast signals should not be run long board edges or pass near connectors.

Gaps in Return Plane - I'd like to come back to the gap or slot in the return plane mentioned earlier and show an example of why it's bad news for EMI. When the return path is interrupted, the conduction current is forced around the slot, or otherwise finds the nearest (lowest impedance) path back to the source. The electromagnetic field is forced out and the field will "leak" all over the board. I have an article and good demonstration video of this and how it affects common mode currents and ultimately, EMI. See *Figure 6* and *Reference 4*.

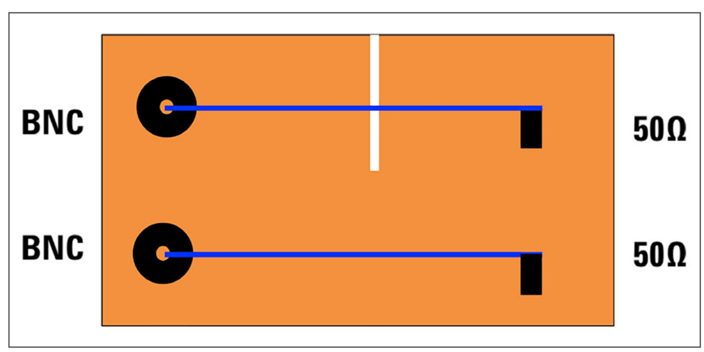

Figure 6 - shows a demonstration test board with transmission lines terminated in 50 Ohms. One transmission line has a gap in the return plane and the other doesn't. A 2 ns pulse generator is connected to one of the two BNC connectors in turn and the harmonic currents in a wire clipped to the return plane are measured with a current probe.

The difference between the gapped and un-gapped traces is shown in *Figure 7*. Note the harmonic currents are 10 to 15 dB higher for the gapped trace (in red). Failing to pay attention to the signal and power return paths is a major cause of radiated emissions failures.

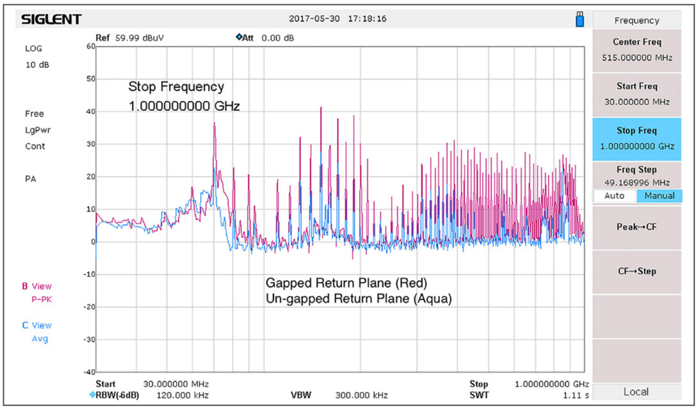

Figure 7 - The resulting common mode currents on an attached wire as measured with a current probe. The trace in aqua is the un-gapped return path and the trace in red, the gapped return path. The difference is 10 to 15 dB higher for the gapped return path. These harmonic currents will tend to radiate and will likely cause radiated emissions failures.

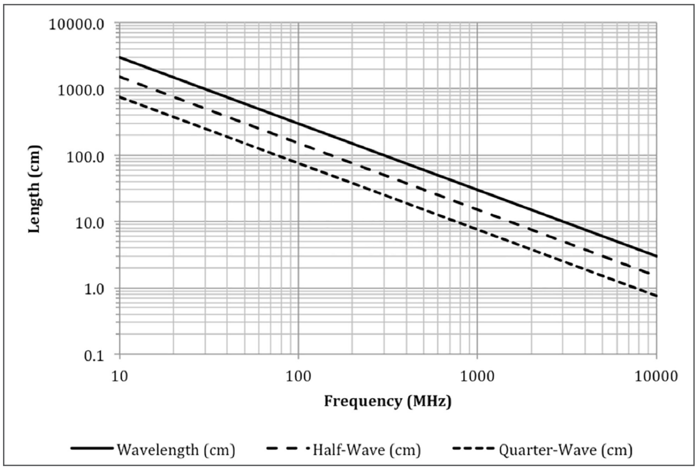

Figure 9 - A handy chart for determining resonant frequency versus cable or slot length in free space. Half-wavelength slots simulate dipole antennas and are particularly troublesome. Figure, courtesy Patrick André.

Shielding

The two issues with shielded enclosures is getting all pieces well-bonded to each other and to allow power or I/O cable to penetrate it without causing leakage of common mode currents. Bonding between sheet metal may require EMI gaskets or other bonding techniques. Slots or apertures in shielded enclosures become issues when the longest dimension approaches a half wavelength. *Figure 8* shows a handy chart for determining the 20 dB attenuation of a given slot length. See *Reference 5* and *6* for more detail on shielding. Interference Technology also has a free downloadable 2016 EMI Shielding Guide with excellent information (*Reference 7*).

Figure 9 is a chart of wavelength versus frequency. For example a 6-inch (15 cm) slot has a half wave resonance at 1000 MHz. If a product design requires at least a 20 dB shielding effectiveness, then the longest slot length can be just one-half inch.

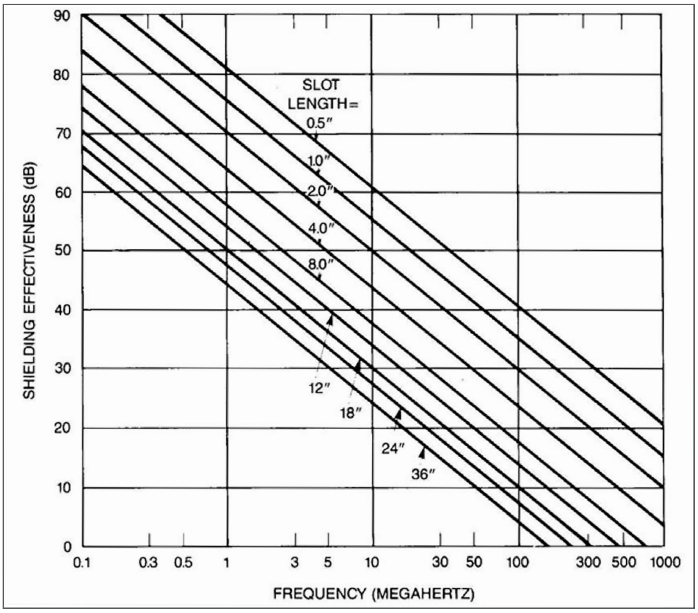

Figure 8 - A chart of attenuation versus slot length. Figure, courtesy Henry Ott.

Cable Penetration - The number one issue I find when tracking down a radiated emissions problem is cable radiation. The reason cables radiate is that they penetrate a shielded enclosure without some sort of treatment - either bonding the cable shield to the metal enclosure or common mode filtering at the I/O or power connector (*Figure 10* and *11*). This occurs frequently, because most connectors are attached directly to the circuit board and are then poked through holes in the shield. Once the cable is plugged in, it is "penetrating the shield" and EMI is the usual result.

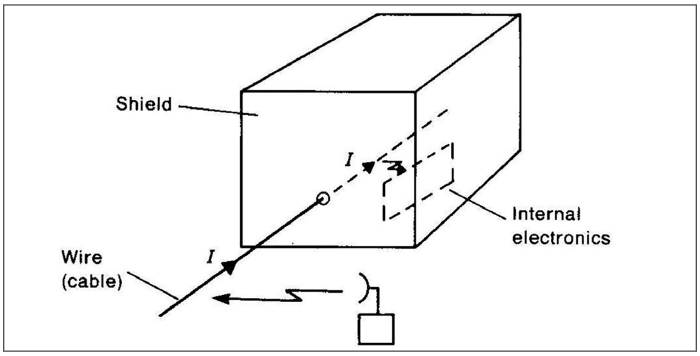

Figure 10 - Penetrating the shield with a cable defeats the shield. This example shows how external energy sources can induce noise currents in I/O cables, which can potentially disrupt internal circuitry. The reverse is also true, where internal noise currents can flow out the cable and cause emissions failures. Figure, courtesy Henry Ott.

There are four combinations or cases that must be considered: shielded or unshielded products, and shielded or unshielded cables. Power cables are usually unshielded for consumer/commercial products and so require power line filtering at the point of penetration or at the connector of the circuit board. Shielded cables must have the shield bonded (ideally in a 360 degree connection) to the product's shielded enclosure. If the product does not have a shielded enclosure, then filtering must be added at the point of penetration or at the I/O connector of the PC board. *Figure 11* shows the usual result when connectors simply poke through a shielded enclosure.

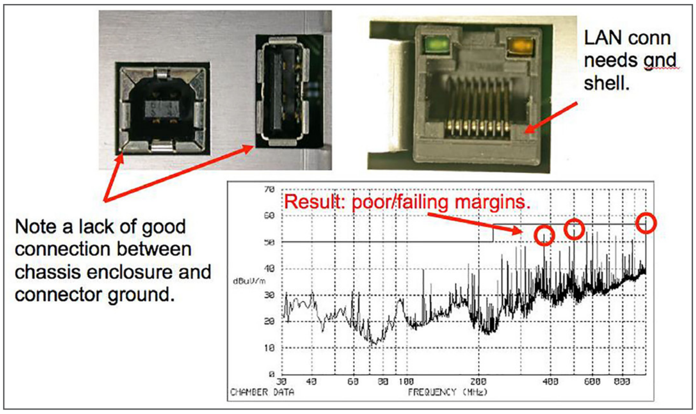

Figure 11 - Result of a penetrating cable through a shielded enclosure, because of un-bonded I/O connectors to the shielded enclosure.

Cable Shield Terminations - Another potential issue is if the I/O cable uses a "pigtail" connection to the connector shell. Ideally, cable shields should be terminated in a 360-degree bond for lowest impedance. Pigtails degrade the cable shield effectiveness by introducing a relatively high impedance. For example, a 1-inch pigtail connection has 12 Ohms impedance at 100 MHz and gets worse the higher you go in frequency. This is especially problematic for HDMI cables, because the HDMI working group (http://www.hdmi.org) failed to specify the method for terminating the cable shield to the connector.

Filtering

I won't go into very much detail here, because Interference Technology has an excellent EMI Filter Guide free for the downloading (see *Reference 8*). Suffice to say, filters, as well as transient protection, are important at power and I/O connectors. Typically, these will be common mode topologies, as shown in *Figure 12*. Most signal-level common mode chokes may be obtained in surface mount packaging. Power chokes are much larger to handle the current and may be obtained as either surface mount or through-hole mount, depending on the current rating. Many Ethernet connectors have built-in common mode filtering.

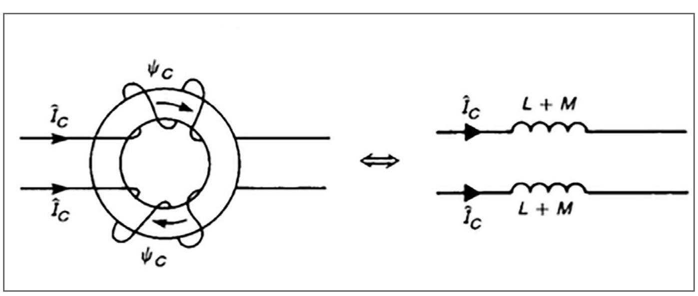

Figure 12 - A typical common mode filter used for I/O filtering. The two windings are wound in opposite directions and so tend to cancel the common mode currents.

Power supply input filters are generally designed to suppress both differential and common mode currents. A typical topology is shown in *Figure 13*. The "X" capacitor is designed to filter differential mode, while the CM choke and "Y" capacitors are designed to filter common mode. The resistor shown is usually 100 kOhm and the purpose is merely to bleed off the line voltage stored on the capacitors to a safe level.

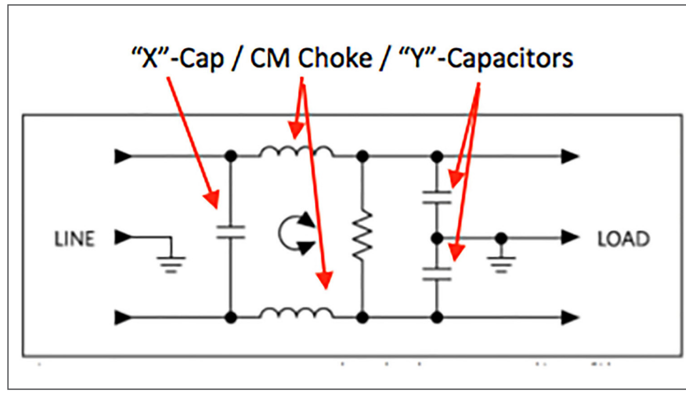

Figure 13 - A general purpose filter typically used for power supply input filtering.

For general purpose filtering of signals, the handy chart of possible filter topologies may be found in *Reference 9* and is reproduced here in *Figure 14*. The appropriate topology depends on the source and load impedances. If these impedances are not known, then either the "PI" or "T" topology may be used (#3 or #5 on the chart, respectively).

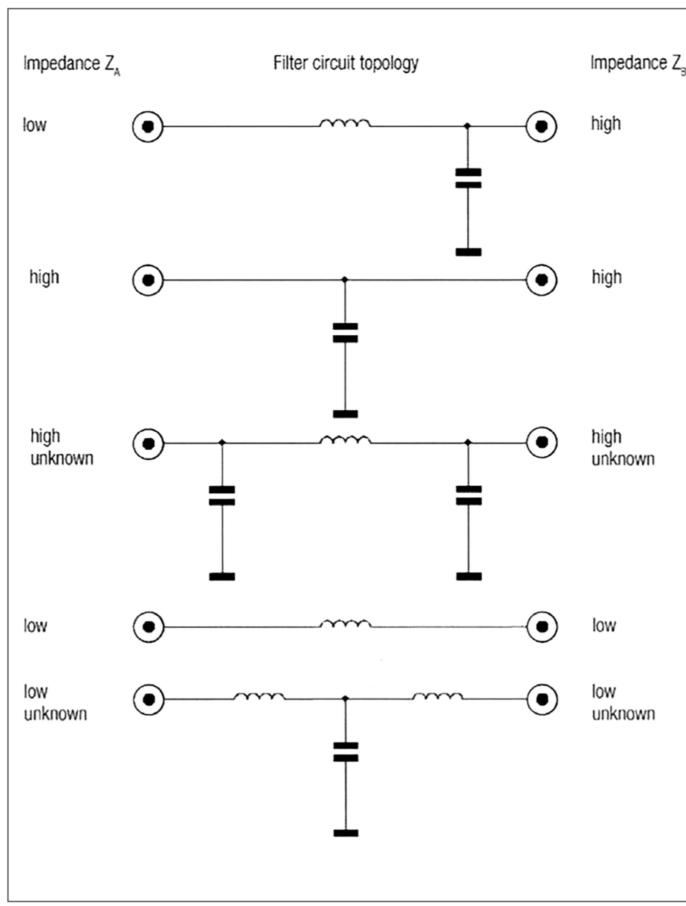

Figure 14 - Five common filter topologies, depending on the source and load impedances. Figure, courtesy Würth Electronik.

Ferrite or inductive components should not be used in series with the power pins of ICs, as this will only reduce the ability of the local decoupling capacitors to supply required energy during simultaneous switching of the IC output stages with the resulting higher power supply noise.

Ferrite Chokes - One common filter element usually added to I/O cables is the ferrite choke. Ferrite chokes come in either the clamp-on types or solid cores meant to be assembled along with the cable assembly. Often, these are used as a last resort to reduce cable emissions or susceptibility. Most ferrite chokes have an associated impedance versus frequency characteristic, often peaking around 100 to 300 MHz. Some materials are designed to peak below 100 MHz for lower frequency applications. Maximum impedances can range from 25 to 1000 Ohms, depending on the ferrite material used and style of choke.

Sometimes, clipping a ferrite choke onto a cable has no effect. This is usually due to the fact the choke has the same, or lower, effective impedance than the cable itself.

The attenuation of a ferrite choke is easily calculated.

Attenuation (dB) = 20 * log((Zin + Zferrite + Zload) / (Zin + Zload))

For example, if we add a 100 Ohm ferrite choke to a power supply cable with system impedance of 10 Ohms, the attenuation would be:

Attenuation = 20 * log((10 + 100 + 10) / (10 + 10)) = 15.5 dB

Refer to *Reference 9* for much additional detail on ferrite chokes and general filter design.

Transient Protection

In order to protect internal circuitry from electrical transients, such as ESD, electrically fast transient (EFT), or power line surge, due to lightning, transient protective devices should be installed at all power and I/O ports. These devices sense the transient and "clamp" the transient pulse to a specified clamp voltage.

Transient protectors in signal lines must generally have a very low parallel capacitance (0.2 to 1 pF, typical) to the return plane, depending on the data rate in order to maintain signal integrity. These silicon-based devices may be purchased in very small surface mount packaging.

Power line surge protection usually requires much larger transient protection devices and they can come in a variety of types. Gas discharge or metal oxide varistors are the most common, but larger silicon-based devices are also available. More information on the design of surge protection may be found in *Reference 4*.

Summary

Most EMC/EMI failures are due to poor shielding, penetration of cables through shields, poor cable shield termination, poor filtering, and above all, poor PC board layout and stack-up. Paying attention to these common design faults will pay off with a lower risk of compliance failures and result in lower project costs and schedule slippage.

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Author Bio

Kenneth Wyatt is president and principal consultant of Wyatt Technical Services LLC, as well as the senior technical editor for Interference Technology Magazine. He has worked in the field of EMC engineering for over 30 years with a specialty is EMI troubleshooting and precompliance testing. He is a co-author of the popular EMC Pocket Guide and RFI Radio Frequency Interference Pocket Guide. He also coauthored the book with Patrick André, EMI Trouble shooting Cookbook for Product Designers, with forward by Henry Ott. He is widely published and authored the EMC Blog hosted by EDN.com for nearly three years. Kenneth is a senior member of the IEEE and a long time member of the EMC Society.

He may be contacted at:
ken@emc-seminars.com
kwyatt@interferencetechnology.com

Check out his web site for more technical information, training schedules, and links:
http://www.emc-seminars.com

TOP THREE EMI AND POWER INTEGRITY PROBLEMS WITH ON-BOARD DC-DC CONVERTERS AND LDO REGULATORS

Kenneth Wyatt
Wyatt Technical Services LLC
ken@emc-seminars.com

Steve Sandler
Picotest Systems, Inc.
steve@picotest.com

Modern devices are continuing a long-term trend of squeezing more electronics into smaller packages, while also increasing system performance, data rates and operating efficiency. Higher efficiencies are often achieved by implementing faster silicon MOSFETs or even faster eGaN FETs while size is reduced by increasing switching frequencies and replacing aluminum and tantalum capacitors with smaller ceramic devices. One result of this trend is that there is greater interaction between the disciplines of EMI, signal integrity (SI) and power integrity (PI).

Introduction

EMI is a measure of the electromagnetic emissions produced by the high-speed current and voltage signals the system creates. Power integrity is a measure of the power quality at the device that being powered. This means that the power supply voltages must be maintained within the allowable operating voltage range of high-speed devices. Devices, such as modems, reference clocks and low noise amplifiers (LNAs) are all sensitive to noise on the power rails, which results in timing jitter, spurious responses reduced data channel eye openings, and degraded signal-to-noise ratio (SNR). This too, is a measure of power integrity. The power supply itself is a noise source and the noise sources generated by the power supply must be kept from propagating through the system.

This article discusses the three most common causes of EMI and power integrity issues while providing tips for how to avoid or minimize them in your design,

1. *Ringing on switched waveforms causes broad resonant peaks in the emission spectrum.*
2. *DC-DC converters generate noise at the switching frequency, and because of high speed switching devices, can generate broadband switching harmonics well into the GHz.*
3. *Power plane resonance in DC-DC converter or LDO regulators due to high-Q capacitors resonating with power planes.*

1. Ringing and Radiated Emissions

Any ringing on the switched waveform (fairly common) can lead to broadband resonances in the resulting RF spectrum. Resonant frequencies resulting from DC-DC converters or low dropout (LDO) linear regulators can be as low as a few kHz while resonance due to the PDN with switching devices, such as MOSFET's can be in hundreds of MHz or higher.

The harmonic energy resulting from this switching is "captured" by the PDN and device resonances, evident as ringing in the time domain. The current and voltage of this ringing produces EMI. The magnitudes of the ringing and EMI are related to the quality factor (Q) and characteristic impedance of the resonance and the harmonic energy produced by the switching.

As an example, the switching waveform on a DC-DC buck converter demo board was measured with a Rohde & Schwarz RTE 1104 oscilloscope and Rohde & Schwarz RT-ZS20 1.5 GHz active probe (*Figure 1*).

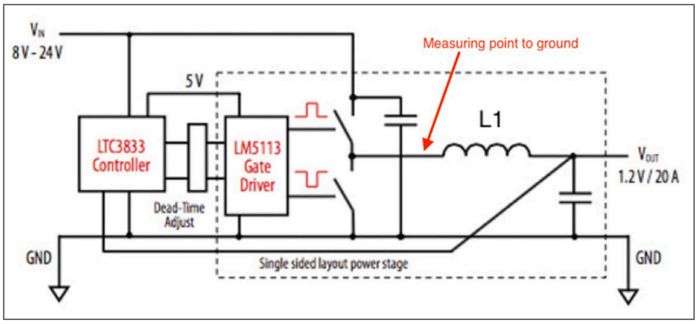

Figure 1. Diagram showing the measuring point at the switch device junction (on the left side of L1) to ground return.

There was a very large ringing superimposed on the switched waveform of 216 MHz. This can be seen clearly in *Figure 2*.

A Fischer Custom Communications F-33-1 current probe was used to measure both the input power cable common mode current (violet trace) and output load differential mode current (aqua trace). See *Figure 3*. Note the broad resonant peaks at 216 MHz (*marker 1*) and the second harmonic at 438 MHz (*marker 2*).

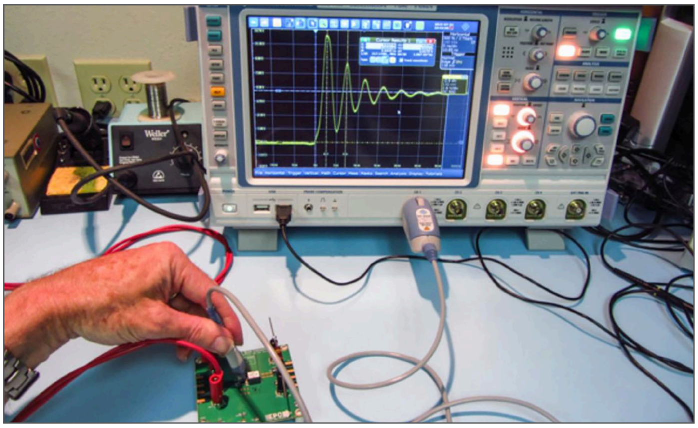

Figure 2. Measuring the rise time and ringing on a DC-DC converter. Notice to strong ringing at 216 MHz.

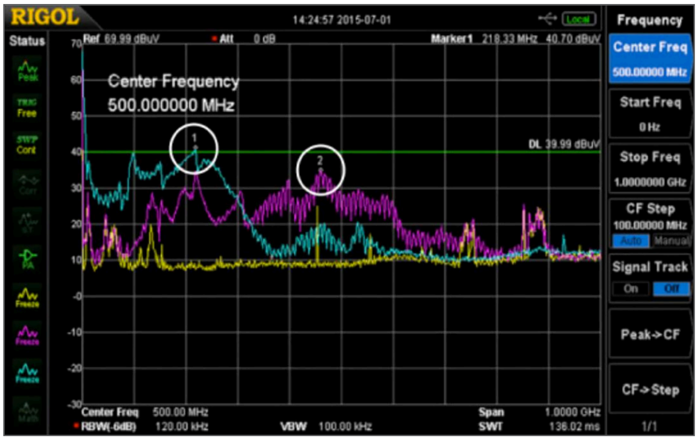

Figure 3. Resulting resonances from the 216 MHz ring frequency (marker 1) and second harmonic at 438 MHz (marker 2).

Remediation Tips - There are several ways to improve the design to minimize the resonances, ringing and therefore EMI. Since the energy is related to the switching frequency, rise time of the switching, characteristic impedance, and Q of the resonances, these factors are also the paths to mitigation.

- Slower edges will degrade operating efficiency but reduce high frequency energy
- Careful PCB design and capacitor selection will minimize the characteristic impedance and Q
- Keep traces short and wide and dielectrics thin.
- Keep all the switching circuitry on one side of the board, preferably with a thin dielectric to the respective ground return plane.
- Use of a snubber circuit, damping of resonances using controlled ESR capacitors, or redesign of the inductor for lower leakage inductance.

For additional detail on measuring ringing refer to *Reference 1*.

2. Fast Edges Create Broadband Noise at GHz Frequencies

Today's on-board DC-DC converters use switching frequencies as high as 3 MHz. This is an advantage because it allows for physically smaller inductor and filter components, as well as increased efficiency. However, the fast edge speeds create broadband harmonic energy. The bandwidth of this harmonic energy is related to the voltage and current rise time. A 1ns edge speed can produce harmonic energy up to 3 GHz, or more.

These broadband harmonics are the cause of radiated emissions failures and also can affect the receiver sensitivity of any on-board telephone modems or other wireless systems, such as GPS. *Figure 4* shows how a typical DC-DC converter circuit can be characterized using an H-field probe connected to a spectrum analyzer.

It's also possible to connect the probe to an oscilloscope and hold it near each DC-DC converter to get some idea of the ringing, if any, without disturbing the circuit.

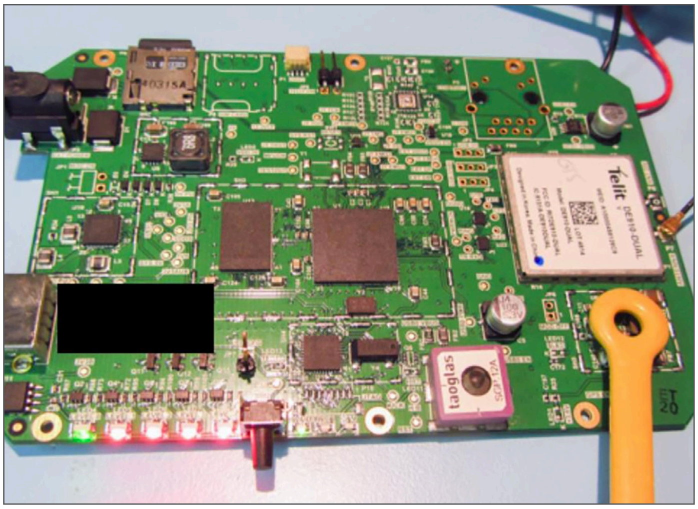

Figure 4. Probing DC-DC converter noise sources on a typical wireless device.

Figure 5 shows the resulting measurement of a couple DC-DC converters. The yellow trace is the ambient noise floor of the measurement system and is always a good idea to record for reference. The aqua and violet traces are the two converter measurements. Note that both produce broadband noise currents out to 1 GHz, with the converter in violet out to beyond 1.5 GHz. Note the violet trace is 20 to 50 dB higher than the ambient noise floor.

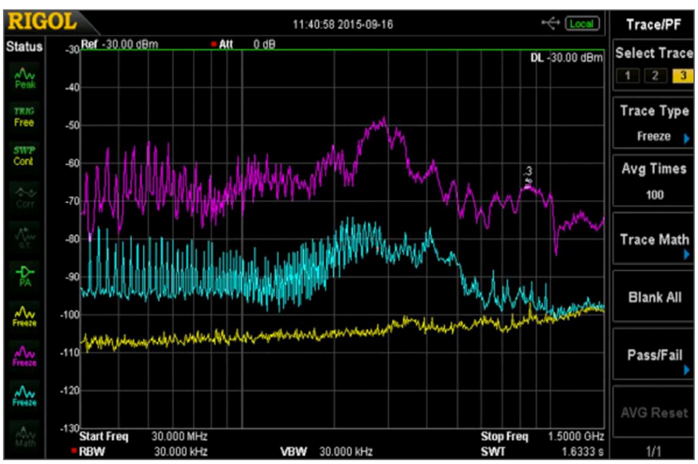

Figure 5 - In this example, we're looking from 30 MHz to 1.5 GHz to generally characterize the spectral emissions profile of a couple of on-board DC-DC converters. Both will potentially cause interference to mobile phone bands in the 700 to 950 MHz region. The one with the violet trace is over 30 dB above the ambient noise level in the mobile phone band.

Remediation Tips – To reduce the risk of self-interference to on-board mobile phone modems and wireless systems, the product design must start off with EMC in mind and with no corners cut. This will consist of:

- A near perfect PC board layout
- Filtering of DC-DC converters
- Filtering of any high frequency device
- Filtering of the radio module
- Local shielding around high noise areas
- Possibly shielding the entire product
- Proper antenna placement

The PC board layout is critical and is where most of your effort should reside. An eight or ten layer stack-up will provide the most flexibility in segregating the power supply, analog, digital, and radio sections and provide multiple ground return planes, which may be stitched together around the board edge to form a Faraday cage. Care must be taken to avoid return current contamination between sections – especially in the ground return planes. For wireless products, the power plane for the radio modem section should be isolated (except via a narrow bridge) from the digital power plane. All traces to this isolated plane should pass over the bridge connecting the two. This can provide up to 40 dB of isolation between the digital circuitry and radio.

It is vital that the power and ground return planes be on adjacent layers and ideally 3-4 mils apart at the most. This will provide the best high frequency bypassing. All signal layers should be adjacent to at least one solid ground return plane. Clock, or other high-speed traces, should avoid passing through vias and should not change reference planes.

Power supply sections should be well isolated from sensitive analog or radio circuitry (including antennas). Be aware of primary and secondary current loops and their return currents. These return currents should not share the same return plane paths as digital, analog, or radio circuits. Remember that high frequency return currents want to return to the source directly under the source trace.

For more details on resolving DC-DC converter noise issues with wireless radio modems, refer to *Reference 2*.

3. PC Board Plane Resonance and the Effect on Radiated Emissions

Noise propagation in a simple system can be represented by three elements, the voltage regulator, the printed circuit board planes with decoupling capacitors (PDN) and the device being powered (load).

Each of these three elements is comprised of resistive, inductive and capacitive terms. Even "noise free" low dropout (LDO) regulators can be highly inductive (*Reference 3*). The resistive, inductive and capacitive terms can resonate amplifying the noise signals created by the power supply and the load as they travel across the PDN creating EMI. The harmonics of the switching frequency and the switch ringing discussed earlier excite these PDN resonances (*Reference 4*). As stated previously this noise can degrade and interfere with on-board wireless modems, as well as resulting radiated and conducted emissions.

A short video helps explain the basic principles of PDN design (*Reference 5*). The radiated EMI of a LTC3880 DC-DC converter measured near the input plane using an H-field probe is seen in *Figure 6*.

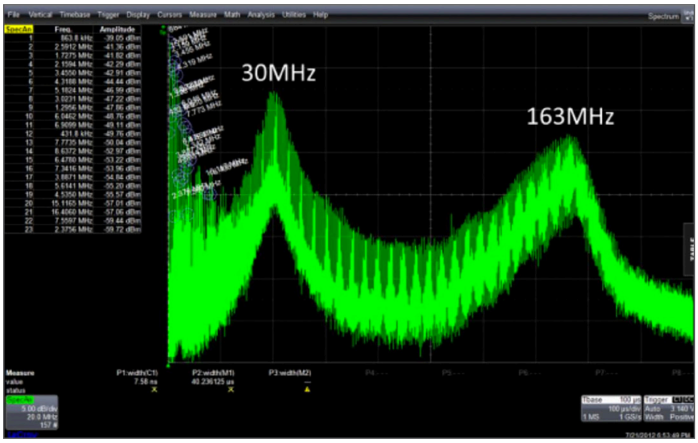

Figure 6. Spectrum analyzer display showing the 30 MHz and 160 MHz resonances detected near the input power connections of a DC-DC converter.

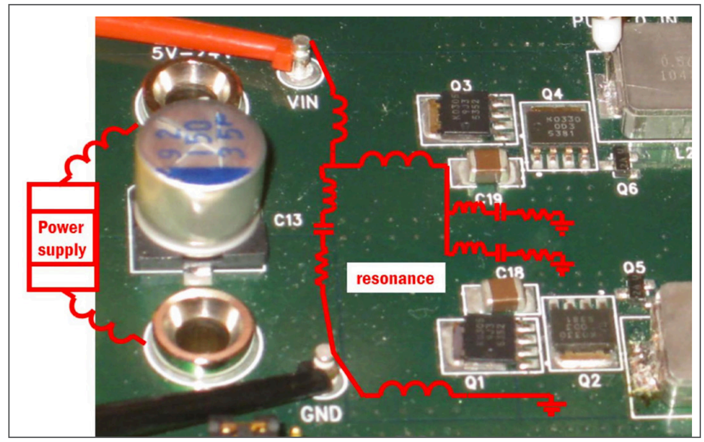

Figure 9. The power plane section of the DC-DC converter (measured in Figure 6) with schematic representations of the component, PC board and external connections.

The 163 MHz is attributed to the ringing of the switches as seen in *Figure 7*. This ringing is caused by the inductance of the upper MOSFET bond wires, pins and circuit board planes, ringing with the lower MOSFET and PC board capacitance.

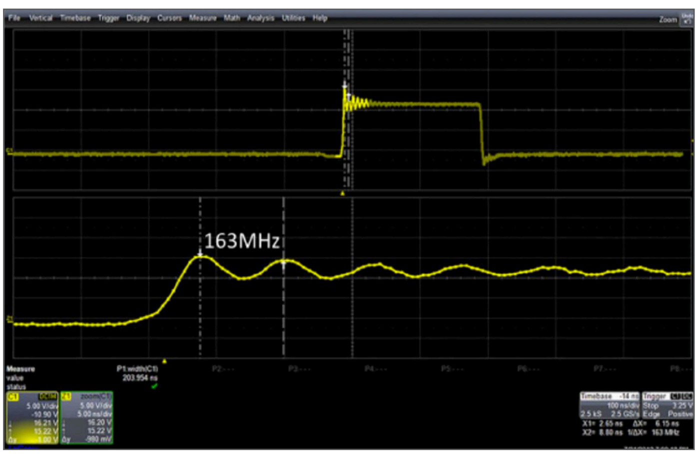

Figure 7. The 163 MHz EMI is easily explained by the ringing at the switch device, as discussed earlier.

The input power plane section of the DC-DC converter (measured in *Figure 6*) is shown in *Figure 9* with schematic representations of the component, PC board and external connections.

A very simple simulation example can be used to illustrate these impedance resonance effects. Consider a simple DC-DC converter as shown in *Figure 10*.

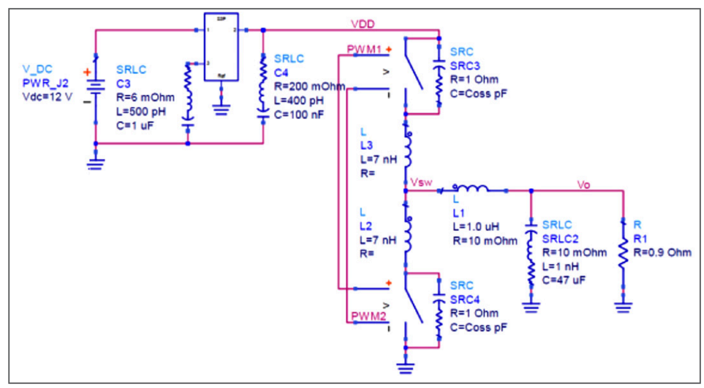

Figure 10. A simple DC-DC converter for illustration of plane resonance EMI. The "FET" switches include lead inductance and drain capacitance (Coss). A small PC board and two ceramic capacitors are included.

The input ceramic decoupling capacitor resonates at approximately 30 MHz, as seen in *Figure 8* and results in the large 30 MHz EMI signature.

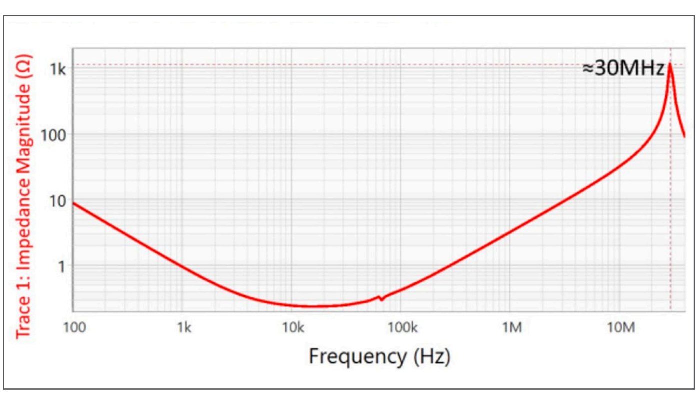

Figure 8. The larger 30 MHz emission is identified as a printed circuit board resonance using an H-field probe and confirmed by a 1-port reflection impedance measurement at the input capacitor.

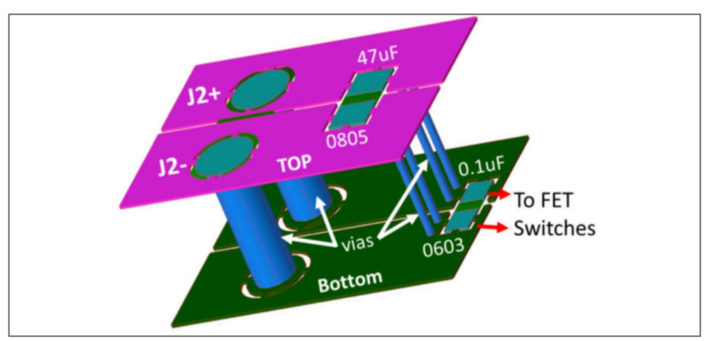

Figure 11. The large round pins on the left are the input power connector, J2. The larger capacitor on the top side is an 0805 sized 47 µF and the smaller capacitor on the bottom side is an 0603 sized 0.1 µF.

Designers frequently place the FET switches on one side of the board with power entry on the opposite side of the PC board. The small PC board plane used in this example has power entry through a pair of pins and no interconnect inductance is added to connect power to the PC board. A large 47 µF ceramic capacitor is placed on the top side of the PC board, while a smaller, 0.1 µF ceramic capacitor is placed very close to the FET switches on the bottom side of the PC board. Two parallel vias connect power and ground from the top side of the PC board to the bottom side as seen in *Figure 11*.

The simple model is used to simulate the harmonic current in the input connector, which is directly related to conducted and radiated emissions. Two simulations are performed; one with low ESR ceramic capacitors and the other with a lower Q controlled ESR ceramic replacing the 0.1 µF capacitor close to the FET switches. Both simulations are shown together in *Figure 12*.

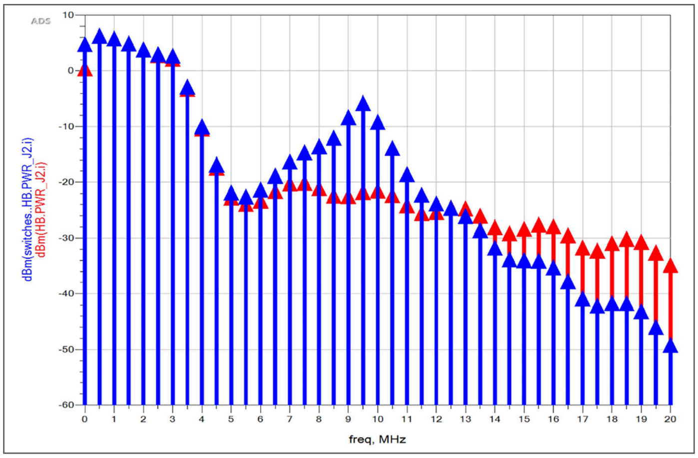

Figure 12. Spectral simulation of the input power lead shows the high Q ceramic (10 mΩ blue) has a clear peak near 10 MHz that is eliminated using a controlled ESR ceramic (200 mΩ red)

The simulated impedance, measured at the smaller capacitor in *Figure 13* shows the corresponding plane resonance with a clear 10 MHz peak using the high Q ceramic capacitor (blue) and the peak is eliminated using the controller ESR ceramic capacitor (red).

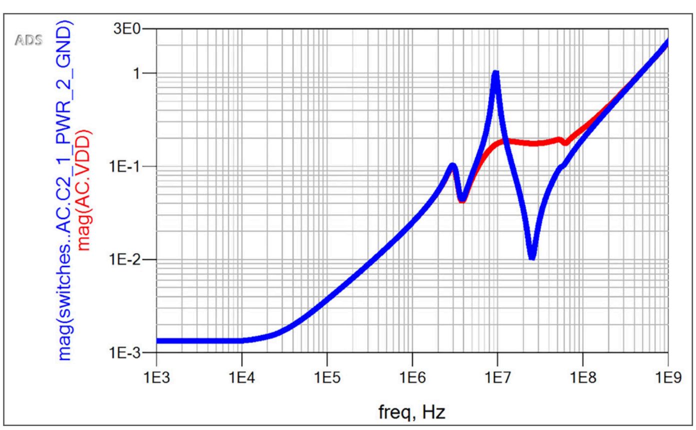

Figure 13. The simulated impedance at the 0.1 uF capacitor using high Q ceramic (10 mΩ blue) and a controlled ESR ceramic (200 mΩ red)

Remediation Tips – To minimize PDN resonances, the complete system of voltage regulator, PDN and the load needs to be carefully balanced. Damping resistance must be included to eliminate or minimize the existence or Q of resonances. This will consist of:

- Short, wide power planes
- Keep the layout as small as possible to minimize inductance
- Thinner PC board dielectric layers, closer to the surface
- Incorporate EM simulation to identify and minimize PDN resonances
- Keep capacitors on one side of the PC board to the extent possible
- Low-Q or ESR controlled capacitors reduce Q
- Choose voltage regulators and output capacitors for good control loop stability
- Don't place cutouts or holes in ground plane layers below the power plane
- Ferrite beads are a very common cause of PDN resonances
- Be aware of inductive interconnects bringing power to the system.

Printed circuit board design and decoupling is critical and "rules-of-thumb" generally don't work well in high speed circuits. The design of the circuit board and capacitor decoupling always involves trade-offs, but the impacts on resonances need to be weighed carefully. A multi-frequency harmonic comb generator can be extremely helpful for quickly identifying PDN resonances (*Reference 3*).

Summary

As you can see, designing DC-DC converters, LDOs, and PDNs with today's high-speed technology nearly always requires careful circuit design, adequate filtering, simulation of the PDN, very careful circuit board layout, and use of controlled-ESR filter capacitors. Poor designs can result in:

- Ringing in power supply switches (or other fast-edged digital switching) resulting in associated radiated or conducted emissions resonant peaks at the ring frequency and harmonics.
- High frequency broadband noise well beyond 1 GHz, resulting in self-interference to radio modems.
- Poor stability and resonances in un-damped power distribution networks, leading to instability, spectral resonances, and associated radiated and conducted emissions.

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BASICS OF PASSIVE FILTERS FOR EMC COMPLIANCE

Don MacArthur
Principal EMC Consultant, MacArthur Compliance Services, LLC
don.macarthur@mcs-emc.com

Introduction
One of the roles of the practicing EMC engineer or product designer is to be able to design filters to add to circuits in order to get them to pass various EMC immunity and emissions standards such as IEC 61000-4-2 for ESD immunity, IEC 61000-4-3 for Radiated RF immunity and IEC 61000-4-4 for Electrical Fast Transient/Burst immunity and other various international standards covering Radiated Emissions (RE) or Conducted Emissions (CE). EMI filters are often used along with proper shielding in order to achieve EMC compliance. The purpose of a filter is to establish either a low-impedance path for RF current to return back to the local source of energy, and/or to provide a high impedance to prevent RF currents from flowing on a cable. However, selecting the proper filter for a given situation may be confusing to some, especially if they are new to the EMC field or have not dealt with the subject in some time. EMC practitioners may be asking themselves what filter configuration is the best one to use for any given application or how to correctly choose the values of components given the frequency, circuit impedance, and other parameters of the circuit. They may also want to know how they can get more attenuation out of their filter design in order to pass an emissions or immunity test. The time to learn how to properly design filters for EMC compliance is not when schedules are tight, and the product's ship date is rapidly approaching. If you find yourself stuck in any of the above situations, this article on passive filter basics for EMC compliance should help remove the mystery, and allow you to quickly find the best passive component filter solution that allows product to ship on time.

Passive Low-Pass Filters

Fortunately, designing filters for EMC compliance is not as difficult as it may seem. For most cases, in order to achieve EMC compliance, we really only need to know how to apply passive low-pass filter types to our circuits. The other types of passive filters, such as high-pass, band-pass, and band-reject are not as common as the low-pass filter is for EMC work and will not be covered in this paper. Consult the references for more information on these other filter types.

Unfortunately, circuit impedances are not always well understood or impossible to know, making it more difficult to determine which values of passive low-pass filter components to choose from in order to pass the EMC compliance tests. This is the situation with common mode emissions emanating off of a cable during a RE test where the impedance of the cable changes as it is rearranged in order to maximize emissions (*Reference [1]*).

It is impossible to model the filter exactly if the load impedance is not known. The only way to know if a low-pass filter design is adequate or not is by trial and error experiments performed during EMC compliance testing, or more preferably, by trying out different low-pass filter component values very early in the product development cycle. In order to be most effective, this experimental work should occur during pre-compliance testing performed in your own test facility prior to going out of house for full-compliance testing. See *Reference [3]* for a detailed description on how to setup an in-house pre-compliance EMC test facility.

A low-pass filter is one in which the frequencies below a certain significant frequency are easily let-through and those above this same significant frequency are heavily attenuated. A passive low-pass filter is a simple voltage divider; non-amplifying device composed of a combination of resistors and capacitors, inductors (or ferrites) and capacitors or in some instances, may be composed of just one of these components. For instance, a single capacitor placed across a line to reference ground without the resistor or inductor installed may be all that is required in order to suppress an unwanted signal.

The benefit to using a single component filter is that only one physical device is required which in turn requires less board space and also helps keep parts costs down. Multi-element filters are useful in situations where the range of frequencies involved is too large and impossible for a one component filter to fully attenuate.

RC Low-Pass Filter

One of the most basic forms of a low-pass filter is comprised of just one resistor and one capacitor, an RC filter. In an RC low-pass filter, the cutoff frequency occurs at resonance, where the capacitive reactance (X_c) equals the resistance (R) and where $X_c = 1/2\pi fC$ (*Reference [4]*).

A simple RC low-pass filter and the equation for determining its cutoff frequency is shown in *Figure 1*. Note that the filter shown in *Figure 1* is also known as an L filter due to its resemblance to the letter L. It is also considered a single-pole filter because there is only one reactive component, the capacitor.

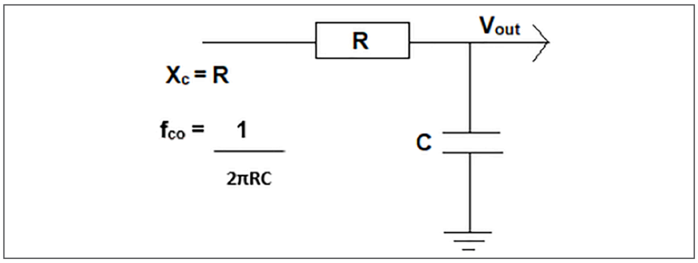


Figure 1: Basic RC Low-Pass Filter (L Type)

A low-pass filter has an ideal, theoretical response where all signals contained below a so-called critical frequency (the 3 dB down point) are easily let-through the device and above which frequency, all signals are heavily attenuated. An ideal low-pass filter response curve is shown in *Figure 2*.

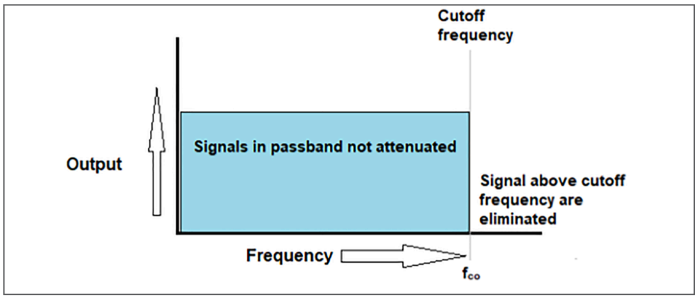


Figure 2: Ideal low-pass filter response curve

In actual practice, the output of the filter will not go to zero as abruptly as shown in the ideal curve of *Figure 2*. In actuality, the output will gradually roll off at a 6 dB/octave or 20 dB/decade rate as shown in *Figure 3*.

EMC Application of Low-Pass Filters

Reference [3] suggests applying a low-pass filter in order to fix an EMC problem such as a fast transient or ESD discharge immunity issue and that a good starting point in putting together a low-pass filter that will work for most situations is to start out by using a 47 to 100Ω series resistor placed in the signal line, with a 1 to 10nF capacitor placed in the signal or power return line. If we take this information and select R = 100Ω and C = 10nF as a starting point, the cut-off frequency (f_{co}) will equal approximately 159 kHz, and the low-pass filter response curve should look like that shown in *Figure 3.* Very little of the signals that are greater than 1.59 MHz will be let through the filter as they are 20 dB lower than any of the signals that at the filter's cutoff frequency of 159 kHz.

As another example, if we leave R = 100Ω and select C = 1nF, the cutoff frequency at the 3 dB down point moves out to roughly 1.59 MHz, the 6 dB down point is at 3.2 MHz, and the signal is almost completely attenuated at 15.9 MHz. Signals greater than 15.9 MHz are heavily attenuated and not let through the filter.

Table 1 contains a matrix of the various R-C low-pass filter values discussed so far plus some others that might be useful, and their low-pass filter characteristic responses at the 6 dB and 20 dB down points.

When attempting to suppress an unwanted high-frequency signal, one may find out that a filter containing only a single reactive component (i.e. one capacitor or one inductor) may not provide enough attenuation. Adding a second reactive component will increase the roll off to 12 dB/octave or 40 dB/decade (*Reference [4]*). These types of filters are called various names such as double-pole, two-stage, two-element, or second-order filters. Filters with three reactive components will provide 18 dB/octave or 60 dB/decade attenuation. Four reactive component filters will provide 24 dB/octave or 80 dB/decade attenuation and so on (*Reference [2]*).

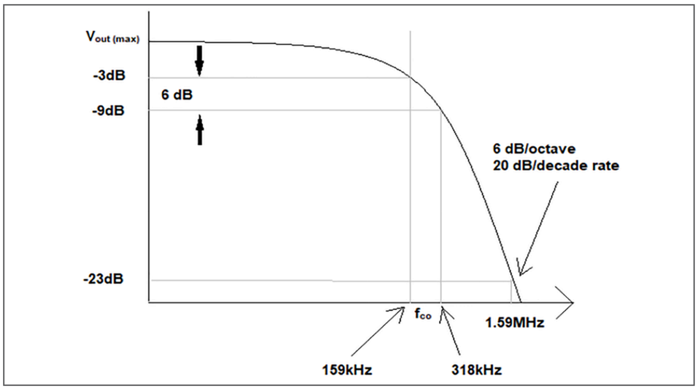


Figure 3: Realistic low-pass filter response curve

R	C	f_{co} (-3 dB Point)	-6 dB Point	-20 dB Point	
200	10nF	79.6 kHz	159.2 kHz	795.8 kHz	
100	10nF	159 kHz	318 kHz	1.59 MHz	← Plotted in Figure 3
49	10nF	325 kHz	650 kHz	3.3 MHz	
20	10nF	796 kHz	1.6 MHz	7.9 MHz	
200	1nF	796 kHz	1.6 MHz	7.9 MHz	
100	1nF	1.59MHz	3.2 MHz	15.9 MHz	
49	1nF	3.3 MHz	6.5 MHz	32.5 MHz	
20	1nF	7.9 MHz	15.9 MHz	79.6 MHz	
200	100pF	7.96 MHz	15.9 MHz	79.6 MHz	
100	100pF	15.9 MHz	31.8 MHz	159.2 MHz	
49	100pF	32.5 MHz	65 MHz	325 MHz	
20	100pF	79.6 MHz	159.2 MHz	796 MHz	

Table 1: Matrix of R-C Values and Low-Pass Filter Reponses

Selection of f_{co}

When selecting a cut-off frequency for a low-pass filter, it is important to take into account the fundamental frequency of the intended data, clocks, and other purposeful signals present on the filtered line. If the cut-off frequency chosen is too low in frequency, then the intended signals will be attenuated along with the higher frequency signals that you want to suppress. Try to maintain at least the 5th harmonic of the intended signal, with the 10th harmonic being ideal (*Reference [3]*). Many I/O signals that are used with unshielded cables require some form of filtering in order to be in compliance with EMC standards. These signals usually have a frequency of 1 MHz or less (*Reference [1]*). It is important to also ensure that by adding a filter's impedance to circuit that it does not in turn create a signal integrity problem.

Once the filter's component values are chosen, carefully consider where it is going to be placed in the circuit or system. The most benefit is obtained when the filter is placed as close to the item to be protected as possible, one centimeter is ideal for most designs (*Reference [1]*). In order to keep any extra unwanted inductance from affecting performance of the filter, be sure to keep lead lengths as short as possible. Additional layout and placement concerns will be covered later in this article.

Use of Ferrites

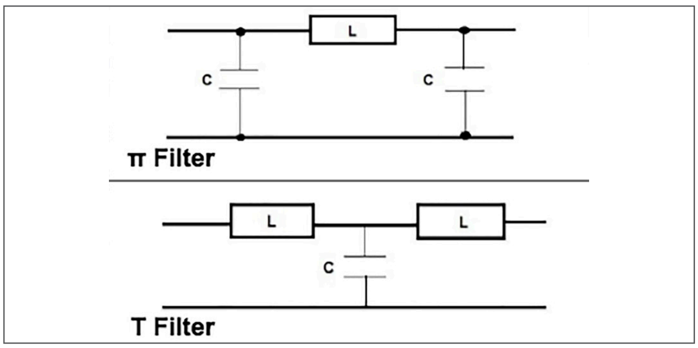
If the voltage drop across the series resistor cannot be tolerated, a device such as a ferrite, which acts as a high-frequency resistor with minimal voltage drop, can be used instead of the resistor. Because the ferrite presents the circuit with high AC impedance, while also not affecting signal quality, they are most optimal for filtering at frequencies greater than 30 MHz. Carefully consider the amount of DC or low-frequency current present in the circuit when using ferrites. They can become easily saturated with too much current present in the circuit which renders them ineffective (*Reference [5]*).

Use of Inductors

An inductor can also be considered for the series element in a low-pass filter instead of a resistor or ferrite, particularly if dealing with a signal in the 10 to 30 MHz range. When using inductors, beware of the effect that their inductive reactance ($X_L = 2\pi fL$) and parasitic capacitance will have at these higher frequencies. You may be actually creating a high-pass filter when you are attempting to create a low-pass one, and not even realize it.

Basic Filter Topologies

The following diagrams show two more of the basic filter configurations available for impedance mismatching between circuit source and load input and output impedances and filter input and output impedances. Both are named after their shapes. The first is called a π filter because it looks like the Greek letter π and the second is called the T filter because it looks like the letter T. Note that there are three reactive elements present in these filters which means they an attenuation curve of 18 dB/octave and 60 dB/decade. They are considered third-order filters (*Reference [5]*).



Impedance Mismatching

Source and load impedances must be considered in selecting the proper filter configuration. If order to work properly, the source driving the input to the low-impedance shunt element (i.e. capacitor), should be a high-impedance. If the output of the source is a low-impedance, it should face the high-impedance series component. This same concept applies to load input impedances versus the filter's output impedances. In general, a source or load impedance less than 100 Ω is considered low and great than 100Ω is considered high impedance (*Reference [5]*). *Table 1* provides a matrix of source versus load impedances and their associated correct filter topologies.

Source Z	Load Z	Filter Configuration	Analysis
High (>100Ω)	High (>100Ω)	Shunt Element (Capacitive) or π Filter	Use π filter if greater roll-off is required.
High (>100Ω)	Low (<100Ω)	L Filter	The shunt element should face the High Z source and this element should face the Z load.
Low (<100Ω)	Low (<100Ω)	Series Element (Inductive) or T Filter	Use T filter if greater roll-off is required.
Low (<100Ω)	High (>100Ω)	L Filter	The shunt element should face the High Z load and the series element should face the Low Z source.

Differential Mode (DM) and Common Mode (CM) Currents
There are two different types of current modes, and hence noise sources capable of creating interference. It is important to know which mode is prevalent so that proper filtering can be applied. The two types of signals we are referring to are differential mode (DM) and common-mode (CM) signals.

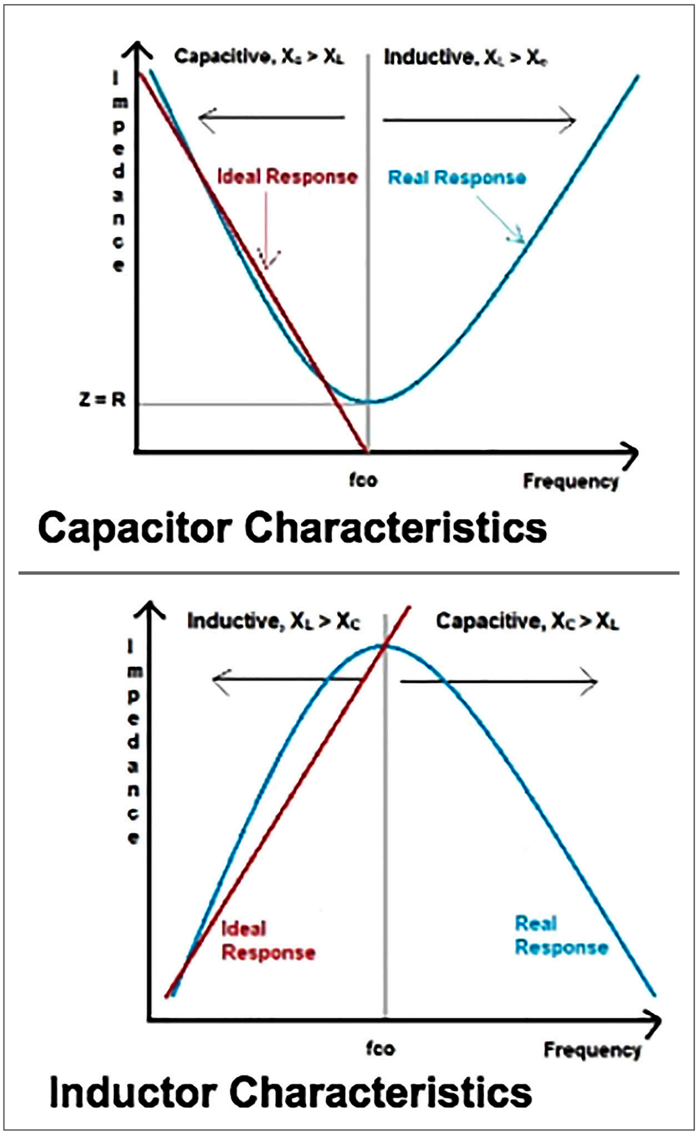
DM signals carry useful information whereas CM currents provide no useful information what-so-ever and are the main source of RE and CE issues. A DM signal travels down one side of a circuit path, and an equal and opposite DM signal travels back on the other side of the path. If no circuit discontinues exists, then complete canceling of these two DM signals occur, and no CM current is developed. Placing capacitors across the outgoing and return lines and/or an inductor in series with either outgoing or return line is called DM filtering.

CM signals are in-phase signals present in both outgoing and return lines of a circuit. They do not cancel each other out but add up, often to a level substantial enough to cause EMI issues. CM filtering involves placing capacitors across each signal line to ground reference and sometimes also using a CM inductor in the circuit. The CM inductor only acts on the CM signals that are present. It does not affect the DM signals.

Parasitics
The non-ideal behavior of the elements that make up our filter must be addressed. Unexpectedly, we will find that real capacitors and inductors possess both capacitance and inductance which limits the bandwidth that they are useful over. The amount of parasitics present in a circuit can be reduced through proper component selection and layout techniques, but cannot be eliminated entirely. As frequency increases, the reactance of a capacitor decreases until it reaches its self-resonant frequency. Up to this point, the capacitor is behaving as it should – it behaves like a resistor. Above its self-resonant frequency point the capacitor becomes inductive and it acts like an inductor because of the parasitic inductance found in its metal plates. This parasitic effect is greater in leaded types of capacitors than it is with the surface mount technology (SMT) types that have almost no lead length.

The opposite effect occurs with an inductor where its reactance becomes capacitive above its self-resonant frequency point, and where the inductor now acts like a capacitor. At the self-resonant frequency, capacitors are intended to provide a very low impedance and inductors should provide a high impedance. For inductors, their limiting factors are related to the parasitic capacitance present between each winding and overall capacitance located between one lead and the other.

The inductor's inter-winding parasitic capacitance is not as big a deal in regards to effectiveness for EMI suppression as is a capacitor's parasitic inductance. The main factors that change the intended behavior of capacitors is the parasitic inductance of the circuits in which they are installed, not necessarily the construction of the capacitor. Therefore, proper layout and placement then becomes the critical factor when attempting to effective utilize passive low-pass filters for EMI suppression.



Layout and Placement Concerns

Because there is going to be unknown and hidden parasitics involved, do not expect your filter design to work one-hundred percent the first time. As mentioned earlier, expect the need to perform some trial and error design and troubleshooting in the lab. If not available already, have on hand a selection of various components that you want to try out. Do not wait until the last minute to obtain the SMT capacitors, inductors, or ferrites that you want to use. Make sure the components selected are designed for the bandwidths involved. Create your own matrix of values, critical frequencies, and 6- and 20-dB roll-off curves.

In reviewing the layout, look for longer than necessary trace lengths that add extra inductance and impedance. When applying fixes, be sure keep connections short. If an R-C filter is added to the reset pin of a micro-controller, place it as close to the pin as possible and do not overlook the length of its return trace. In general, it is best to locate the filter as close to the offending signal source as possible, not some obscure location far away.

Watch out for trace or wire routing that allows for too much capacitive and inductive coupling to other noisy signal or traces. Filter components should be placed right at an entry connector (I/O and power inputs). Placement of a filter deeper inside a circuit or system allows EMI to enter the system (*Reference [6]*). If separation is not maintained, improper routing of input and output sections can mean that filter elements are essentially bypassed and no longer effective. On PCBs, capacitors should shunt unwanted signals to chassis not line to line or line to return (*Reference [6]*). It is best to understand the path of current flow and to not necessarily rely on "ground" as being the ultimate zero-ohm impedance and sole problem savior.

Finally, although they appear to be useful and easy to troubleshoot with, do not expect too much out of clamp-on ferrite common-mode chokes as they only provide about 10 dB of attenuation (*Reference [3]*).

Conclusion

The need to utilize passive low-pass filters to obtain EMC compliance is a given. They provide a low-impedance path for RF currents to return back to the local source of energy or provide a high impedance to prevent unwanted RF currents from flowing. A filter that does both is ideal. Designing low-pass filters for EMI suppression is not that difficult. Proper knowledge and planning before the need for them arises can save developers some time and headaches.

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SUMMARY OF MILITARY AND AEROSPACE EMC TESTS

Ghery Pettit
Pettit EMC Consulting
Ghery@PettitEMCConsulting.com

Introduction

Military and aerospace EMC tests cover a wide range of products. While the standards, including limits and test methods may differ, all EMC test standards have a few things in common. The most basic are the limits for emissions and the types and levels of susceptibility testing.

Emissions tests (and their associated limits) are put in place for military and aerospace equipment primarily to protect other systems from interference. These other systems may or may not include radio equipment. Examples abound showing the effect of inadequate EMC design. The Interference Technology 2016 Military EMC Guide (Reference 1) provides 3 such examples on page 11.

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Download your copy at:
http://learn.interferencetechnology.com/2017-military-and-aerospace-emc-guide/

To download the 2018 edition, visit:
https://learn.interferencetechnology.com/2018-military-and-aerospace-emc-guide/

While many military and aerospace EMC issues may be addressed by operational changes, testing is still required to find weaknesses.

Military and aerospace EMC testing is performed at the system and subsystem levels. MIL-STD-464C provides requirements at the system or platform level. The latest version, MIL-STD-461G, provides requirements at the equipment or subsystem level. *Reference 1* provides details on both of the standards, but this article will highlight some key tests, particularly as they relate to MIL-STD-461G.

RATIO	DESCRIPTION
CE101	Conducted Emissions, Audio Frequency Currents, Power Leads
CE102	Conducted Emissions, Radio Frequency Potentials, Power Leads
CE106	Conducted Emissions, Antenna Port
CS101	Conducted Susceptibility, Power Leads
CS103	Conducted Susceptibility, Antenna Port, Intermodulation
CS104	Conducted Susceptibility, Antenna Port, Rejection of Undesired Signals
CS105	Conducted Susceptibility, Antenna Port, Cross-Modulation
CS109	Conducted Susceptibility, Structure Current
CS114	Conducted Susceptibility, Bulk Cable Injection
CS115	Conducted Susceptibility, Bulk Cable Injection, Impulse Excitation
CS116	Conducted Susceptibility, Damped Sinusoidal Transients, Cables and Power Leads
CS117	Conducted Susceptibility, Lightning Induced Transients, Cables and Power Leads
CS118	Conducted Susceptibility, Personnel Borne Electrostatic Discharge
RE101	Radiated Emissions, Magnetic Field
RE102	Radiated Emissions, Electric Field
RE103	Radiated Emissions, Antenna Spurious and Harmonic Outputs
RS101	Radiated Susceptibility, Magnetic Field
RS103	Radiated Susceptibility, Electric Field
RS105	Radiated Susceptibility, Transient Electromagnetic Field

Table 1: MIL-STD-461G Emission and Susceptibility Requirements

MIL-STD-461G divides test requirements into 4 basic types. Conducted Emissions (CE), Conducted Susceptibility (CS), Radiated Emissions (RE) and Radiated Susceptibility (RS). There are a number of tests in each category and *Table I*, taken from MIL-STD-461G Table IV, shows these test methods.

A brief description of each these tests will be provided below. These are summarized from a more detailed introduction to MIL-STD-461G, which is found in the *References 1, 2*, and *3*. Keep in mind that a complete copy of MIL-STD-461G is 280 pages, so any information here is brief and the standard must be read and understood. A copy of MIL-STD-461G may be obtained free. See *Reference 4*.

CE101 Conducted Emissions, Audio Frequency Currents, Power Leads. CE101 is applicable from 30 Hz to 10 kHz for leads that obtain power from sources that are not part of the EUT. There is no requirement on output leads from power sources. Emission levels are determined by measuring the current present on each power lead. There is different intent behind this test based on the usage of equipment and the military service involved. The specific limits are based on application, input voltage, frequency, power and current.

CE102 Conducted Emissions, Radio Frequency Potentials, Power Leads. CE102 is applicable from 10 kHz to 10 MHz for leads that obtain power from sources that are not part of the EUT. There is no requirement on output leads from power sources. The lower frequency portion is to ensure EUT does not corrupt the power quality (allowable voltage distortion) on platform power buses. Voltage distortion is the basis for power quality so CE102 limit is in terms of voltage. The emission levels are determined by measuring voltage present at the output port of the LISN. Unlike CE101, CE102 limits are based on voltage. The basic limit is relaxed for increasing source voltages, but independent of current. Failure to meet the CE102 limits can often be traced to switching regulators and their harmonics.

CE106 Conducted Emissions, Antenna Port. CE106 is applicable from as low as 10 kHz to as high as 40 GHz (depending on the operating frequency) for antenna terminals of transmitters, receivers, and amplifiers and is designed to protect receivers on and off the platform from being degraded by antenna radiation from the EUT. CE106 is not applicable for permanently mounted antennas.

CS101 Conducted Susceptibility, Power Leads. CS101 is applicable from 30 Hz to 150 kHz for equipment and subsystem AC and DC power input leads. For DC powered equipment, CS101 is required over the entire 30 Hz to 150 kHz range. For AC powered equipment, CS101 is only required from the second harmonic of the equipment power frequency (120 Hz for 60 Hz equipment) to 150 kHz. In general, CS101 is not required for AC powered equipment when the current draw is greater than 30 amps per phase. The exception is when the equipment operates at 150 kHz or less and has an operating sensitivity of 1 µV or better.

The intent is to ensure that performance is not degraded from ripple voltages on power source waveforms.

CS103, CS104 and CS105 Conducted Susceptibility, Antenna Port, Intermodulation, Rejection of Undesired Signals and Cross-Modulation. This series of receiver front-end tests include test methods for Intermodulation (CS103), Rejection of Undesired Signals (CS104) and Cross Modulation (CS105). They were designed for traditional tunable super-heterodyne type radio receivers. Due to the wide diversity of radio frequency subsystem designs being developed, the applicability of this type of requirement and appropriate limits need to be determined for each procurement. Also, requirements need to be specified that are consistent with the signal processing characteristics of the subsystem and the particular test procedures to be used to verify the requirement.

CS109 Conducted Susceptibility, Structure Current. CS109 is a highly specialized test applicable from 60 Hz to 100 kHz for very sensitive Navy shipboard equipment (1 µV or better) such as tuned receivers operating over the frequency range of the test. Handheld equipment is exempt from CS109. The intent is to ensure that equipment does not respond to magnetic fields caused by currents flowing in platform structure. The limit is derived from operational problems due to current conducted on equipment cabinets and laboratory measurements of response characteristics of selected receivers.

CS114 Conducted Susceptibility, Bulk Cable Injection. CS114 is applicable from 10 kHz to 200 MHz for all electrical cables interfacing with the EUT enclosures.

CS115 Conducted Susceptibility, Bulk Cable Injection, Impulse Excitation. CS115 is applicable to all electrical cables interfacing with EUT enclosures. The primary concern is to protect equipment from fast rise and fall time transients that may be present due to platform switching operations and external transient environments such as lightning and electromagnetic pulse.

CS116 Conducted Susceptibility, Damped Sinusoidal Transients, Cables and Power Leads. CS116 is applicable to electrical cables interfacing with each EUT enclosure and also on each power lead. The concept is to simulate electrical current and voltage waveforms occurring in platforms from excitation of natural resonances with a control damped sine waveform.

CS117 Conducted Susceptibility, Lightning Induced Transients, Cables and Power Leads. CS117 is one of two new test methods added to MIL-STD-461G. CS117 is applicable to safety-critical equipment interfacing cables and also on each power lead. Applicability for surface ship equipment is limited to equipment located above deck or which includes interconnecting cables, which are routed above deck. The concept is to address the equipment-level indirect effects of lightning as outlined in MIL-STD-464 and it is not intended to address direct effects or nearby lightning strikes.

CS118 Conducted Susceptibility, Personnel Borne Electrostatic Discharge. CS118 is applicable to electrical, electronic, and electromechanical subsystems and equipment that have a man-machine interface. It should be noted that CS118 is not applicable to ordnance items. The concept is to simulate ESD caused by human contact and test points are chosen based on most likely human contact locations. Multiple test locations are based on points and surfaces which are easily accessible to operators during normal operations. Typical test points would be keyboard areas, switches, knobs, indicators, and connector shells as well as on each surface of the EUT.

RE101 Radiated Emissions, Magnetic Field. RE101 is applicable from 30 Hz to 100 kHz and is used to identify radiated emissions from equipment and subsystem enclosures, including electrical cable interfaces. RE101 is a specialized requirement, intended to control magnetic fields for applications where equipment is present in the installation, which is potentially sensitive to magnetic induction at lower frequencies.

RE102 Radiated Emissions, Electric Field. RE102 is applicable from 10 kHz to 18 GHz and is used to identify radiated emissions from the EUT and associated cables. It is intended to protect sensitive receivers from interference coupled through the antennas associated with the receiver.

RE103 Radiated Emissions, Antenna Spurious and Harmonic Outputs. RE103 may be used as an alternative for CE106 when testing transmitters with their intended antennas. CE106 should be used whenever possible. However, for systems using active antenna or when the antenna is not removable or the transmit power is too high, RE103 should be invoked. RE103 is applicable and essentially identical to CE106 for transmitters in the transmit mode in terms of frequency ranges and amplitude limits. The frequency range of test is based on the EUT operating frequency.

RS101 Radiated Susceptibility, Magnetic Field RS101 is a specialized test applicable from 30 Hz to 100 kHz for Army and Navy ground equipment having a minesweeping or mine detection capability, for Navy ships and submarines, that have an operating frequency of 100 kHz or less and an operating sensitivity of 1 µV or better (such as 0.5 µV), for Navy aircraft equipment installed on ASW capable aircraft, and external equipment on aircraft that are capable of being launched by electromagnetic launch systems. The requirement is not applicable for electromagnetic coupling via antennas. RS101 is intended to ensure that performance of equipment susceptible to low frequency magnetic fields is not degraded.

RS103 Radiated Susceptibility, Electric Field. RS103 is applicable from 2 MHz to 18 GHz in general, but the upper frequency can be as high as 40 GHz if specified by the pro-

Equipment and Subsystems Installed In, On, or Launched From the Following Platforms or Installations	TYPE OF PRODUCT/SERVICE																		
	CE101	CE102	CE106	CS101	CS103	CS104	CS105	CS109	CS114	CS115	CS116	CS117	CS118	RE101	RE102	RE103	RS101	RS103	RS105
Surface Ships	A	A	L	A	S	L	S	L	A	S	A	L	S	A	A	L	L	A	L
Submarines	A	A	L	A	S	L	S	L	A	S	L	S	S	A	A	L	L	A	L
Aircraft, Army, Including Flight Line	A	A	L	A	S	S	S		A	A	A	L	A	A	A	L	A	A	L
Aircraft, Navy	L	A	L	A	S	S	S		A	A	A	L	A	L	A	L	L	A	L
Aircraft, Air Force		A	L	A	S	S	S		A	A	A	L	A		A	L		A	
Space Systems, Including Launch Vehicles		A	L	A	S	S	S		A	A	A	L	A		A	L		A	
Ground Army		A	L	A	S	S	S		A	A	A	S	A		A	L	L	A	
Ground Navy		A	L	A	S	S	S		A	A	A	S	A		A	L	L	A	L
Ground, Air Force		A	L	A	S	S	S		A	A	A		A		A	L		A	

Table Legend:
A: Applicable (green) L: Limited as specified in the individual sections of this standard. (yellow) S: Procuring activity must specify in procurement documentation. (red)

Table 2: MIL-STD-461G Requirement matrix

curing agency. It is applicable to both the EUT enclosures and EUT associated cabling. The primary concern is to ensure that equipment will operate without degradation in the presence of electromagnetic fields generated by antenna transmissions both onboard and external to the platform. The limits are platform dependent and are based on levels expected to be encountered during the service life of the equipment. It should be noted that RS103 may not necessarily be the worst case environment to which the equipment may be exposed.

RS105 Radiated Susceptibility, Transient Electromagnetic Field. RS105 is intended to demonstrate the ability of the EUT to withstand the fast rise time, free-field transient environment of EMP. RS105 applies for equipment enclosures which are directly exposed to the incident field outside of the platform structure or for equipment inside poorly shielded or unshielded platforms and the electrical interface cabling should be protected in shielded conduit.

Not all tests are required for each type of device or intended use environment. MIL-STD-461G provides a matrix in Table V showing how these tests are used based on the intended use of the device.

Again, the reader is referred to *References 1* through *3* for more details, or to MIL-STD-461G for the details of the standard (*Reference 4*). This guide also provides a list of standards that apply to various military equipment.

A popular and common aerospace EMC requirement required by the FAA for commercial aircraft is RTCA/DO-160, Environmental Conditions and Test Procedures for Airborne Equipment. The latest version is RTCA/DO-160 G, published on December 8, 2010, with Change 1 published on December 16, 2015. DO-160 covers far more than just EMC issues, but the EMC subjects covered include input power conducted emissions and susceptibility, transients, drop-outs and hold-up; voltage spikes to determine whether equipment can withstand the effects of voltage spikes arriving at the equipment on its power leads, either AC or DC; audio frequency conducted susceptibility to determine whether the equipment will accept frequency components of a magnitude normally expected when the equipment is installed in the A/C; induced signal susceptibility to determine whether the equipment interconnect circuit configuration will accept a level of induced voltages caused by the installation environment; RF emissions and susceptibility; lightning susceptibility; and electrostatic discharge susceptibility.

This document can be purchased from RTCA on their website (*Reference 5*). A manufacturer producing products subject to the requirements in RTCA/DO-160 should obtain a copy and ensure they have a complete understanding of the content of the document and that any laboratory testing to it is properly accredited.

Examples of differences in test equipment between commercial and military standards.

There are differences in test equipment used compared with commercial EMC tests. Some examples are provided below.

Where 50 μH LISNs are universally required for commercial EMC tests, there are specific cases for CE01 and CE02 tests where a 5 μH LISN is called out. Limits for CE101 tests are provided in dBμA. LISNs are only used for line impedance stabilization. The measurements are taken with current probes. Limits for CE102, on the other hand, are given in dBμV and measurements are taken in much the same way as for commercial standards with the receiver connected to the RF output port of one of the LISNs and the other RF output port(s) terminated in 50 Ohms. It should be noted that MIL-STD-461G calls out a 20 dB pad on the output of the LISN to protect the receiver from transients. This is not a requirement in the commercial standards, but is worth considering when setting up a laboratory for commercial testing, as well.

Military EMC standards, such as MIL-STD-461G will require the use of different antennas for radiated emissions testing. Commercial equipment standards, such as CISPR 32 and ANSI C63.4, require the use of linearly polarized antennas and do not contain requirements for magnetic field testing.

MIL-STD-461G, RE101, requires the use of a 13.3 cm loop sensor, not required in the commercial standards. A receiver capable of tuning from 30 Hz to 100 kHz is needed.

MIL-STD-461G, RE102, requires testing of radiated emissions to as low as 10 kHz. From 10 kHz to 30 MHz a 104 cm (41 inch) rod antenna is used. This frequency range is not covered in CISPR 32 or the FCC Rules for radiated emissions. Thus, the antenna and receiver requirements are different. From 30 MHz to 200 MHz a biconical antenna is used, also commonly used in commercial testing. From 200 MHz to 1 GHz a double ridge horn antenna is called out in 461G. This is different than the tuned dipole or log periodic dipole array antennas used for commercial testing.

The test procedures are also different for radiated emissions testing, requiring different laboratory set-ups and test facility types. No turntable is needed for MIL-STD-461G, nor is an antenna mast capable of moving the antenna over a range of heights.

MIL-STD-461G, RS103, can require significantly higher field intensities for radiated susceptibility testing. Where CISPR 35 requires 3 V/m from 80 MHz to 1 GHz and at a few discrete frequencies up to 5 GHz (with the option of testing a few discrete frequencies at up to 30 V/m), MIL-STD-461G requires testing from 20 V/m to as high as 200 V/m over the range of 2 MHz to 40 GHz for certain equipment. Additional test equipment (signal generators, amplifiers, antennas, etc.) is required over that needed for commercial testing.

Each test in MIL-STD-461G requires its own unique test equipment. Some may be useable for commercial testing, others may not. If testing to MIL-STD-461G, ensure that the equipment is proper for the tests being performed. A detailed understanding of the requirements in MIL-STD-461G is required to ensure that the proper equipment is being used and the laboratory is following the appropriate processes.

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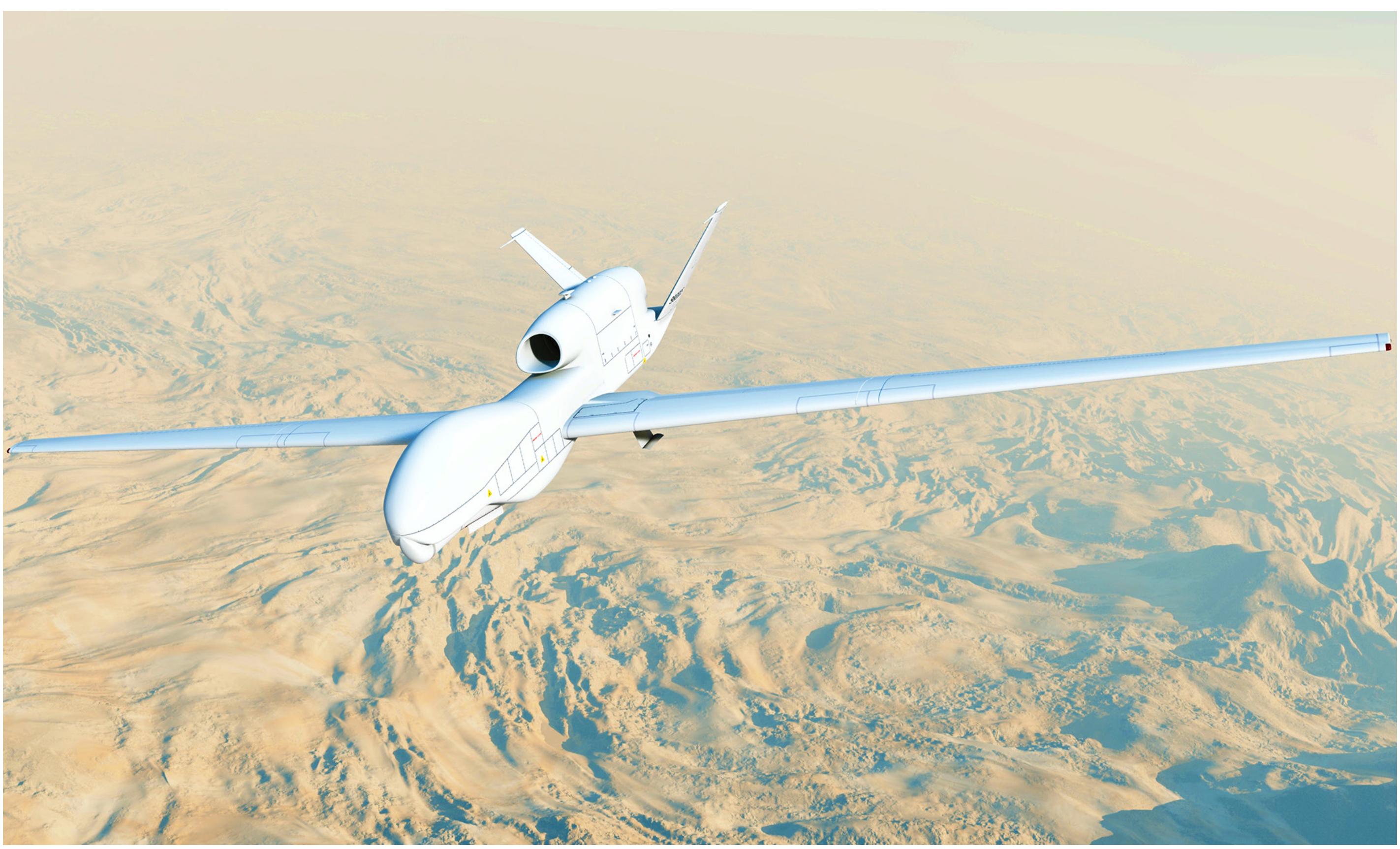

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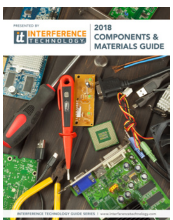

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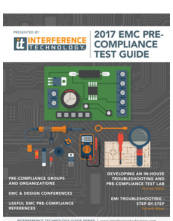

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