• Senior FAE supporting automotive Tier-1 customers throughout Germany
• Over two decades of experience managing EMI challenges
• Deeply involved in the definition and compliance testing of our leading AEC-Q100 power management solutions.
• 22 years at Linear Technology
  o Strategic Marketing Manager for Europe - Product definition and product support for PSU and LED circuits
  o Field Application Engineer
• Additional:
  o Design Engineer, Quality Assurance, Materials Engineer
• Microelectronics. Dipl. Ing., Elektrotechnik University of Dortmund
Automotive EMI Deep Dive Seminar - Agenda

- DC/DC Converter Review
- EMI Coupling Mechanisms
- Magnetic Coupling & Demo Videos
- Layout Analysis & Hints
- Inductor EMI (Non Isolated)
- Isolated Layout EMI and Hints
Identify The Hot Loop

The switches S1 and S2 are alternating “on”. This creates a rectangular voltage at SW which bumps between Vin and GND.
A second order low pass filter at the SW node we make our DC Voltage
Identify The Hot Loop

With the low pass filter

Vout is now a nice DC
The area of voltage * time1 in the “on” period is the same area voltage * time2 in the “off” period.

Both areas are equal
Identify The Hot Loop

Let's look what the current in the inductor does.

The inductor current goes up when S1 is closed and goes down when S2 is closed.
Identify The Hot Loop

Let's look what the current in the inductor does.

The inductor current goes up when S1 is closed and goes down when S2 is closed.
Identify The Hot Loop

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Let's look what the current in the inductor does.

The inductor current goes up when S1 is closed and goes down when S2 is closed.
Identify The Hot Loop

We assume that the current from the input supply and away from the Cout capacitor is DC and has no “ripple”.
Identify The Hot Loop

So we can reduce the EMI circuit analysis to this.
We know that the current in the inductor has that triangular shape.
Identify The Hot Loop

So that triangular current will flow here
In S1 we have a full switched current which moves to and from zero very fast.
Identify The Hot Loop

S2 is fully switched too
Identify The Hot Loop

So we have a high $\frac{dl}{dt}$ loop here
We call it the Hot Loop

Identify The Hot Loop
Identify The Hot Loop

High $dl/dt$ components are the $C_{in}$ and $S_1, S_2$

$S_1, S_2$ can be inside an IC or discrete components
The bottom switch can be replaced with a diode, the high \( \frac{dl}{dt} \) loop keeps the same.
A buck can be operated “backwards” with feeding current into the inductor. It is now a Boost.
On a boost the Hot Loop are now the switches S1, S2 and Cout
Identify The Hot Loop

You can replace the top switch with a diode, the Hot Loop remains the same.
Identify The Hot Loop

Hot Loop of a SEPIC
Identify The Hot Loop

Hot Loop of a Inverting topology

Inverting

-Vout

Vin

Hot loop
Identify The Hot Loop

4 switch buck boost

Vin

Hot loop

Hot loop

Vout
Identify The Hot Loop

Isolated Offline Flyback

High HF Impedance

1

2

10mH

3

4
## MLCC Class II, III Dielectrics

For Class II, III, and IV Ceramic Dielectrics

<table>
<thead>
<tr>
<th>First Character</th>
<th>Second Character</th>
<th>Third Character</th>
</tr>
</thead>
<tbody>
<tr>
<td>Alpha Symbol</td>
<td>Low Temp</td>
<td>Numeric Symbol</td>
</tr>
<tr>
<td>Z</td>
<td>+10°C</td>
<td>2</td>
</tr>
<tr>
<td>Y</td>
<td>-30°C</td>
<td>4</td>
</tr>
<tr>
<td>X</td>
<td>-55°C</td>
<td>5</td>
</tr>
<tr>
<td></td>
<td></td>
<td>6</td>
</tr>
<tr>
<td></td>
<td></td>
<td>7</td>
</tr>
<tr>
<td></td>
<td></td>
<td>8</td>
</tr>
<tr>
<td></td>
<td></td>
<td>9</td>
</tr>
</tbody>
</table>

Material from TDK
Ceramic Capacitor Construction

\[
C = \varepsilon \varepsilon_r \frac{A}{d}
\]

C : CAPACITANCE
K : CONSTANT
n : # OF LAYERS
S : OVER LAPPING AREA
t : LAYER THICKNESS

1. DEVELOPMENT HI-K MATERIAL: K \(\rightarrow\) LARGE
2. LAMINATION TECHNOLOGY

- **2.1 THIN LAYER : t \(\rightarrow\) SMALL**
  - (THICKER)
  - (THINNER)
  - \(t > t'\)

- **2.2 HI # OF LAYERS : n \(\rightarrow\) LARGE**
  - 40 \(\rightarrow\) 400 LAYERS

- **2.3 ACCURATE LAMINATION : S \(\rightarrow\) LARGE**
  - DIFFERENCE
  - DIFFERENCE \(\rightarrow\) SMALL
  - DIFFERENCE
ESL, ESL And Break Down

ESL Values (Generic) by Case Size

<table>
<thead>
<tr>
<th>Case EIAJ</th>
<th>Case EIA</th>
<th>ESL Value (in µH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0603</td>
<td>0201</td>
<td>250</td>
</tr>
<tr>
<td>0816</td>
<td>0306</td>
<td>300-350</td>
</tr>
<tr>
<td>1005</td>
<td>0402</td>
<td>300</td>
</tr>
<tr>
<td>1220</td>
<td>0508</td>
<td>350-400</td>
</tr>
<tr>
<td>1608</td>
<td>0603</td>
<td>450</td>
</tr>
<tr>
<td>1632</td>
<td>0612</td>
<td>250-300</td>
</tr>
<tr>
<td>2012</td>
<td>0805</td>
<td>700-800</td>
</tr>
<tr>
<td>3216</td>
<td>1206</td>
<td>1100-1300</td>
</tr>
<tr>
<td>3225</td>
<td>1210</td>
<td>1000-1200</td>
</tr>
<tr>
<td>4532</td>
<td>1812</td>
<td>1400-1600</td>
</tr>
<tr>
<td>5750</td>
<td>2220</td>
<td>1700-1900</td>
</tr>
</tbody>
</table>

Note: ESL varies slightly by capacitance value within same case size

Breakdown Voltage

The voltage — proof level of ceramic is very high and is strong also on surge voltage

1206 1µF 16V to 1206 3.3µF 10V

Info from TDK
Capacity Vs. Voltage And Temp

**Typical DC Bias Curve (room temp)**

-25% @ 3.3V → 7.5µF

- 0603 10µF X5R 6.3V

**Info from TDK**

- εr of BaTiO3 is f (E)

<table>
<thead>
<tr>
<th>Class Type</th>
<th>Capacitance</th>
<th>Frequency</th>
<th>Voltage</th>
</tr>
</thead>
<tbody>
<tr>
<td>Class I</td>
<td>1000 pF and lower</td>
<td>1.0 MHz</td>
<td>±0.2 Vrms</td>
</tr>
<tr>
<td></td>
<td>Over 1000 pF</td>
<td>1.0 kHz</td>
<td>±0.2 Vrms</td>
</tr>
<tr>
<td>Class II</td>
<td>10 µF and lower</td>
<td>1.0 kHz</td>
<td>±0.2 Vrms</td>
</tr>
<tr>
<td></td>
<td>Over 10 µF</td>
<td>120 Hz</td>
<td>±0.2 Vrms</td>
</tr>
</tbody>
</table>

Not realistic in most applications

Taiyo Yuden

0805 10µF X7R 10V

6µF @ 3.3V 5µF @ 5V
DC Bias vs. Voltage Rating

1µF X5R and X7R from 0402 to 1210

1µF 0402 6.3V
1µF 1210 50V
1µF 1206 50V

Info from TDK
Capacity Vs. Voltage Different $\varepsilon R$

DC Bias Characteristic by Dielectric $K \equiv \varepsilon_r$

Item: Three items by $K$ with 1.8um
Data is average of 3 items

$K \equiv \varepsilon_r$
Capacity Vs. Voltage Different Thickness

DC Bias Characteristic by Dielectric K

Item: Three items by layer thickness with $K=3500$ material
Data is average of 3 items

Graph showing the change in capacitance ($\Delta C/C$) as a percentage vs. voltage (V) for different thicknesses: 1.5um, 1.6um, and 1.8um designs.
Measure A Biased Capacitor With A Simple LCR Meter

\[ V \rightarrow \text{DUT} \rightarrow N \times 1000 \mu F \rightarrow V \]
Measure A Biased Capacitor With A Simple LCR Meter

![Capacitor Circuit Diagram]

- **Capmeter**
- **V**
- **DUT**
- **N*1000uF**
Capacity vs. Time 22uf 1210 X7R biased To 3V9

22uF X7R 1210 3V9 Bias at 1Khz
Capacity over Time
Capacity vs. Time 22uf 1210 X7R biased To 3V9

22uF X7R 1210 3V9 Bias at 1Khz
Capacity over Time

Capacitor Value [uF] vs. Time [h]
Capacitor Impedance

All 0402 capacitors wind up with similar inductance. Even if they are inductive, the impedance is still low and they work as bypass elements.

With TDK SEAT 2013
Capacitor Impedance

The inductance comes only from the case magnetic field volume integrated to infinity.

The 0402 MLCC have an inductance as a 0402 copper block.
EMI Coupling Mechanisms

Automotive EMI Deep Dive Seminar

June, 2018
Designer think in circuit operation with schematics using ideal components
Added Component Parasitics

With components parasitic’s but still concentrated components
Added PCB Resistance

With PCB resistance added

- ESR
- ESL
- C

[Diagram of a circuit with components labeled ESR, ESL, and C, along with the text "With PCB resistance added"]
Additional Parasitics

With inductive, capacitive and resistance of PCB added

Subcircuit added for all nodes
E-Field / Capacitive Near-Field Coupling

With E-Field/capacitive near-field coupling of components and PCB added
H-Field / Magnetic Near-Field Coupling

With H-Field/magnetic near-field coupling and shielding of components and PCB added
Far-Field Radiation

With electromagnetic far-field radiation and reception added
Don’t Give Up.

There is hope.

Not all effects are relevant in all components.

We will concentrate on the dominant effects first for non isolated DC-DC converters.
Capacitor Magnetic Coupling

From an EMI view, the bigger problem of capacitors is that they couple near-field magnetic like two transformer windings. That’s a wideband (GHz) coupling path and defeats their filter function.
Recap: E-Field / H-Field

E-Fields are created by (AC) voltages.

H-Fields or magnetic fields are created by currents.

Current always flows in circles / closed loops – they never end in ground planes or a “nail in the flower pot.”

So hunt for the full circle.
Capacitor Layout Connections

Layout parasitic inductance reduces the effectiveness of capacitors.

**Figure 11-25.** Inductance of various 0805 SMT decoupling capacitor mounting configurations.

Decoupling Capacitor Layout Considerations

- Avoid switching sides
- Avoid using any traces at all for critical decoupling capacitors.
- Use the shortest and most massive layer fill.
- Avoid thermal traps for critical capacitors. You can get away without tombstoning if the thermal layer mass on both sides is comparable or large.
Magnetic Coupling & Demo Videos

Automotive EMI Deep Dive Seminar

June, 2018
Example – MPQ4430 Buck Converter EVB Modified to Show Capacitor Coupling

- Standard EVB

- Modified EVB

- C1A removed

- Only one large input bypass capacitor C1B

- Inductor L2 placed on the back side to eliminate its field
MPQ4430 Buck Converter EVB 4-Layer PCB

Layout of this 4-layer board

Top layer

Inner 1 layer

Inner 2 layer

Bottom layer
MPQ4430 Buck Top layer detail
MPQ4430 Buck Converter Schematic
0805 MLCC Capacitor Probe

Pick-up formed like what a capacitor will see on the board.
Buck Converter Capacitor Magnetic Coupling to 30MHz

Probe
Switcher VIN bypass Cap
Two output caps

12.1080V 3.3570V
0.056mA 0.0000A
0.001 W Eff= 0.4
Buck Converter Capacitor Magnetic Coupling to 30MHz
Buck Converter Capacitor Magnetic Coupling to 300MHz
Why is the magnetic field amplitude and bandwidth so much higher on the input bypass caps?
Harmonics of triangular waveforms decay for higher frequency with $N^2$. $N$ = harmonic number.
In S1 we have a full switched current which moves to and from zero very fast.
Synchronous Buck Converter – Low Side

Harmonics of rectangular waveforms decay for higher frequency only with N

S2 is fully switched too
So we have a high \( \frac{dl}{dt} \) loop here

We call it the “Hot Loop”
Magnetic Antennas

Every trace on your PCB transports AC current

If you create AC current, it forms a magnetic dipole antenna

That radiates in the same way that an electric dipole does

Radiation goes up proportional to area and current
Buck Converter

Harmonics of rectangular waveforms decay for higher frequency only with $N$.

The harmonics of a rectangular waveform go down over frequency as a first-order low pass.

Induction law voltage $\sim \frac{dB}{dt}$ responsible for the magnetic coupling increasing with frequency makes a first-order high pass.

Combined, it is responsible for the surprisingly flat frequency response of magnetic coupled EMI.

By increasing the size of a bypass MLCC or adding additional MLCC for improvement, you often only increase the size of the magnetic pickup and increase the EMI instead of reducing it.
Skin And Proximity Effect

High frequency current density of a rectangular conductor on a single layer PCB
Skin And Proximity Effect

As a general rule, conductors fight internal magnetic fields.

That means no current density internal in the conductor.

Currents are forced to the outside boundary or skin.

In its extreme form this occurs as quantum effect in a super conductor down to DC and is called London depth (nm to um).

For ordinary conductors we call it skin depth
Skin And Proximity Effect
Skin And Proximity Effect

Shielding effect of conductive planes

Return current flow concentrates here
Skin And Proximity Effect

Skin effect isolates return current on opposite sites
Eddy Currents

Here we have our cancelling dipoles
Magnetic Antennas

For any given aspect ratio, cutting \([d]\) in half reduces the radiated field by 6dB

Pictured: PCB cross-section, side-view
Eddy Current Demonstration (Video)
Eddy Currents – Cancellation Effect

Eddy currents can cancel the original magnetic excitation field

They do it better when the eddy current conducting plane and the original magnetic excitation field are closer

Cancellation works better with:

**Low profile components** (flat, short and wide)

**A solid ground plane** without gaps and holes under and near the excitation field
Magnetic Antennas

On AC current loops like hot loops, there is still an increased current density at the outside boundary of the loop.
Magnetic Antennas

That’s the reason you might cut even the GND of the hot loop on the top side. Otherwise there will still be some current density left at the edge of your PCB where it will radiate, and eddy current shielding is less effective.
Vias

Complete through via.
Inductance ~1nH
But what makes it worse, it makes holes into your eddy current shielding layers.
Vias

Thermal image of eddy currents that a rectangular loop produces in a conducting plane
Now eddy current cut with a 4 via row
Vias

Similar with a linear cut
Vias

With 4 block of 4 vias
With a line of 4 vias, but with enough distance that there is still copper in between the vias.
Vias – What Is Bad?
Vias – What Is Bad?

Cuts in GND plane!
Vias connecting into the next plane leave its eddy current conductivity intact.

Yes

No holes
Planes – How Do They Compare?

Magnetic field sensor

0dB single layer board

-11.3dB with 2mm plane distance typ. dual layer board

-30.4dB with 0.2mm plane distance typ. multilayer board
Passive antennas have the same transmit and receive properties. Transformer coupling magnetic fields are attenuated at the transmitting and receiving site by a solid plane.

Magnetic field sensor

0dB single layer board

-46.4db with 2mm plane distance typ. dual layer board

-86.3db with 0.2mm plane distance typ. multilayer board
Planes – How Do They Compare?

PCB components rising high above the plane are the most prone parts to high frequency transformer coupling.
Planes – How Do They Compare?

The GND plane eddy current will attenuate the transmitting as well as the receiving component – so the attenuation multiplies.

The GND plane eddy currents attenuates the emitter as well as the receiver in dB
Planes – How Do They Compare?

It's an aspect ratio game.

The flatter the components, the lower the resulting magnetic field.
Inductors sorted for their radiation  Same conditions

Tested in a GTEM cell with only the inductor inside the GTEM, with a buck switcher under the same Vin, Vout, Iout and switching frequency conditions.
Layout Analysis & Hints

Automotive EMI Deep Dive Seminar

June, 2018
Layer Stack

Use a 4-layer or more board if possible

Spacing between L1 + L2 and L3 + L4 as close as possible. 50µm is perfect (<250µm target)

DC/DC with local GND in L2

Placement and routing
Power + DC/DC-GND
GND
Routing

Local GND area for cooling

• Use a local GND for the Buck-Cin and Diode on component site.
• Make sure the SW-loop is small!
• Place EMI filter components close to connector.
• Do not let the noise from the DC/DC enter the cable!
Layer Stack High Power

Use a 4-layer or more board if possible

Spacing between L1 + L2 and LN-1 + LN
As close as possible
50µm is perfect
<250µm target

DC/DC with local GND in L2

Placement and routing
Power + DC/DC-GND
GND
Quiet GND for filter etc.

• Use a local GND for the Buck-Cin and Diode on component site.
• Make sure the SW-loop is small!
• Place EMI filter components close to connector.
• Do not let the noise from the DC/DC enter the cable!
First identify what type of switcher:

Non isolated DC-DC in a non-symmetric application -> Layout hints 1

Isolated DC-DC with symmetric input lines like a off-line AC-DC -> Layout hints 2
Layout Type 1:

Identify switcher topology like

Buck, Boost, SEPIC, Inverting/BUK, 4-Switch Buck-Boost

Identify the hot loop – see appendix C
Layout Type 1: Hot Loop Buck
Make the hot loop inner area as small as possible and place a solid plane without cuts (no vias other than GND vias) in the next layer under the hot loop with as least dielectric distance as your PCB manufacturer allows.

Use the lowest profile decoupling capacitors (MLCC). As a first order estimate for its magnetic field, you can assume that the hot loop current flows through the middle of the part. That gives you an idea of the elevation over the next layer solid plane and its effectiveness to cancel out the magnetic AC field with eddy currents.
Place the hot loop components and its driver stages as far away from the supply cable/connector and EMI input filter area.

If possible, place input filter on the other PCB side vs switcher.

If spacing is too tight, use all additional shielding available like top heat sink up to a total shielding enclosure for the switcher.
Do not pull decoupling currents into the filter. The connection between the switch-mode VIN decoupling and the EMI filter should always be inductive.
Use a multilayer PCB. It’s easily over 20dB better than a two layer PCB.
Place small capacitors (0402) directly at the output terminal to avoid that your load cable radiates.

This creates problems >100MHz due to inductor parasitic capacitance leakage.

If your layer stack allows it, make a quiet GND layer at the bottom side where all the filter components are grounded, but not the hot loop area.
Layout Type 1: PCB

- Start always with the best PCB stack and multilayer PCB with enough planes that you do not have to make compromises on shielding/eddy current carrying planes under the hot loop and EMI filters.
- If you figure that you have enough EMI margin left you can make more compromised designs to save money in a second step.
- If you get a severely compromised layout, don’t even try to fix it with bandaids like copper foil, ferrite beads, additional filter capacitors etc – it is wasted time.
- All what you likely manage is to get a ballooning effect. You improve the emissions in one small area and make it much worse in several other areas.
- Start over again with a clean layout and go from there. Even multilayers on a fast track do cost less than your wage between breakfast and lunch.
Layout Type 1: EMI Debugging

- Make a clean layout utilizing all former tips.
- Debugging a rotten layout is a waste of everybody's time.
- Use a magnetic field probe and assure that the area around the input filter is quiet. Otherwise you have transformer coupled noise into your filter and no filter component will work as expected other than as a wide band RF coupler. Adding capacitors in such a condition will increase the EMI noise.
- Place a ferrite clamp over the input line. If you get a wideband reduction in conducted and radiated emissions then most likely your inductor and switch node radiates E-Field.
- Place a E field shield over the inductor and switch node to verify this.
- If emissions are still too high, your load cables might radiate. Use a very small load (light bulb over Cout), or decouple the load with ferrite clamps. Place small (0402) capacitors direct at the output terminals. If that does not help your GND is polluted with di/dt currents.
Layout Type 1: EMI Debugging

• If an E-Field shield over the inductor (easy to test with a soldered-on copper foil) will pass the EMI limits but project constraints do not allow for it, you have a few options:
  • **Choose a flat inductor.** Place capacitors (one side grounded, usually Vout caps) on both sides close to the inductor. Orientation of the inductor is important. The outer winding should connect to the quiet node.
  • **Place pins that are higher than the inductor on both sides.** The idea is to reduce the E-Field coupling to the universe.
  • **Place a single or several wires over the inductor**, or any PCB with a solid plane. If you have a design with stacked PCBs use them as E-Field shield over the inductor/switch node.
  • If all falls short, use a single or two-layer small PCB plugged above the inductor/switch node as E-Field shield.
• 4th order filter with 2 inductors or smaller are more effective than a single inductor filter.
• Choose inductors to attenuate the fundamental and the first few harmonics.
• Higher frequency attenuation (>30MHz) completely depends on parasitic magnetic and E-field coupling.
• Connection to the Switching VIN decoupling need to be an inductor.
• Use magnetic cancelling techniques for capacitor placement.
• Inductors should be as low profile as possible.
Common electronic design practice is to make a savvy design, and in a next step reduce it for economics reasons that fulfills customer demand on the market.

There was in the 1950 area a famous businessman which made his fortune from cutting parts out of TV sets until they did not work anymore, and then he replaced the last cut out part and shipped the unit to save cost. It was called Muntzing. [https://en.wikipedia.org/wiki/Muntzing](https://en.wikipedia.org/wiki/Muntzing)

There is a nice story from Bob Pease about Muntzing [http://www.electronicdesign.com/boards/whats-all-muntzing-stuff-anyhow](http://www.electronicdesign.com/boards/whats-all-muntzing-stuff-anyhow)

But if it comes to EMI, the industry practice is even worse. Instead of starting with a savvy design with good chances to pass EMI tests, designs are done that barely have their primary function but no thought, energy and money is dedicated to pass EMI. Then some poor old gray hair chaps at the end of the long overdue project schedule has to throw in all sorts of expensive band aids to get the unit though EMI. Resulting in costly redesigns, external filter material etc.

EMI debugging on a rotten layout with no working planes and quiet spots to hook any filters and test-shielding makes you go in endless circles…

Sure every project managers promises that next time…. 
With all the pain from EMI, some companies try to come up with computer aided EMI design solutions.

On circuit design, SPICE and all its derivate are well known and indispensable in areas where you can’t easily clip out or solder in some components or even measure a signal. Why not use this technique and enhance it enough that it covers EMI?

Above a handful of MHz, coupling effects other than the ideal SPICE connections dominate the EMI behavior. One problem is the high attenuation from the switch node energy which need easy over 120dB damping. Even if you would have accurate models numeric noise/problems will show up with such high dynamic.

But the real problem is not that much the E- and H-field solver which would be too slow anyway. It’s the models of components which will become extremely complex if they are of any use. Try today to get a decent diode recovery simulation from SPICE.

Even if the modelling is solved you would need a quantum computer to get results within the engineers life- or better project-time.

I rather take the quantum computer which is already available. It’s the real application. It gives you any day honest answers if you dare to ask.

And often the problem sits in front of the screen instead in the probed circuit....
Layout Type 1: Identify the Hot Loop

Identify the hot loop and make it as small in diameter as possible.

For example, in a non-synchronous buck place the freewheeling diode and the input capacitor close to each other both pointing in the IC direction and share a common small GND area.
Layout Type 1: Solid Ground Plane

Place a solid plane under the hot loop, filters etc.
Place your input filter as far away from magnetic and E-field radiators like the hot loop. For this reason we put the EMI input filter often on the back side.

If you can't move it as far away from the input bypass caps and place a “quiet” GND plane under the filter, 4th order EMI filters (2 inductors) work out smaller than 2nd order filters with only one larger inductor. We should know by now why physical smaller lower profile components work better…
Layout Type 1: Radiated And Conducted EMI

Up to a few MHz circuits behave as in their SPICE simulation. So if you hunt an EMI problem on the fundamental or first few harmonics you can still think as SPICE with concentrated components.

Above a few MHz for all EMI measures magnetic field and electric field coupling between components and PCB traces dominates. So the EMI measures above a few MHz are the same for radiated and conducted emissions.
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

470kHz with 4u7 XEL6060
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Inner Layer 3
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Inner Layer 4
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Monopole

All measurements done at 13V5 in and 3V3 12A out
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Monopole E-Field PK fixed frequency

150kHz

30MHz
Fixed frequency switcher has the same spurs in PK, QP and AV.

Noise level is 10dB lower in AV

150kHz

Monopole E-Field AV fixed frequency

30MHz
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

150kHz
Monopole E-Field AV spread spectrum
30MHz
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Monopole E-Field AV spread spectrum with top heat sink as shield
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Radiated 30MHz to 1GHz

All measurements done at 13V5 in and 3V3 12A out
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Radiated PK 30MHz to 1GHz fixed frequency
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

30MHz Radiated AV 30MHz to 1GHz fixed frequency 1GHz
Radiated AV 30MHz to 1GHz spread spectrum
Radiated AV 30MHz to 1GHz spread spectrum with top heat sink as shield
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Conducted 150kHz to 30MHz

All measurements done at 13V5 in and 3V3 12A out
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Conducted fixed frequency

150kHz

108MHz
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Conducted spread spectrum

150kHz

108MHz
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Conducted spread spectrum with top heat sink as shield

150kHz - 108MHz
Layout Type 1: Example A 12A Two Phase Sync Design 2x 4436

Conducted fixed frequency with top heat sink as shield

150kHz

108MHz
There is very little frequency dependent effect in the reduction. For filter component dependent you would assume 1th or 2\textsuperscript{nd} order filter functions

Main EMI coupling/radiating effects are:

- Capacitive = E-field coupling/radiating
- Magnetic = Transformer coupling mainly between capacitors
- Galvanic conductive = Voltage created by direct $Z \times I$ (ground bounce)

For the first, you need fast $dU/dt$ present on the switch node

For the latter two, you need $di/dt$ currents as present in all hot loop components and traces
Layout Type 1: EMI Is Called Black Magic

Because:

You can’t locate any of the main EMI creating effects in the schematic.

Capacitors and other components rising above PCB plane level couple magnetic like transformer windings.

Inductors couple/radiate mainly with their E-field/capacitive. More as higher they rise above PCB plane level. They couple with their magnetic field too but that’s often the lesser problem

High di/dt current currents create voltage in planes and traces (bounce).
Layout Type 1: MPQ4430 Demo Board Monopole

Sweep Graph

limits: 0dBm to 50dBm
Transducer: 2MHz
Trace: AV

150kHz to 30MHz
Layout Type 1: MPQ4430 Demo Board Monopole Inductor Rotated 180 Degrees

Sweep Graph

150kHz

30MHz
Layout Type 1: MPQ4430 Demo Board Conducted

Sweep Graph

AV

150kHz

108MHz
Layout Type 1: MPQ4430 Demo Board Conducted Inductor Rotated 180 Degrees

Sweep Graph

150kHz

108MHz
Layout Type 1: MPQ4430 Demo Board Radiated

Sweep Graph

AV

30MHz

1GHz

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Layout Type 1: MPQ4430 Demo Board Radiated Inductor Rotated 180 Degrees
Layout Type 1: Cuts in GND

If in doubt, don’t
No cuts for secondary switcher’s like POL etc. only the battery supplied switchers have an input harness to radiate.

- Make labyrinth cuts if you plan to cut more than one layer. Then the next layers can carry eddy currents
- Do not line up cuts because RF current density is highest on the edges and will magnetic couple into the next planes
Inductor EMI (Non Isolated)

Automotive EMI Deep Dive Seminar

June, 2018
CM Currents Inductor And SW Node

On most EMI measurements the DUT is hooked on a harness

Typical conducted and radiated setup
CM Currents Differential And Common Mode

Differential Mode current
Common Mode current
Where do the CM and DM currents flow?
CM Currents Differential And Common Mode

We can reduce the CM currents to a single wire
We can reduce the CM currents to a single wire
CM Currents Inductor And SW Node

That reduces this setup to this

For RF experts this is a capacitive top loaded T-antenna.

PCB

CM current

50 Ohm LISN
CM Currents Inductor And SW Node

If the only RF energy source is on the PCB: the only way that a CM current flows is when a capacitive or E-field transducer on the PCB is present.

Any loop current (magnetic source) on the PCB cannot create a CM current because magnetic fields are closed loop rotational fields. Think about how an RF net flux through the wire harness should be created if you only have closed loop currents on the PCB.
To create a CM current on the PCB with in a single trunk wire harness there must be an E-Field radiator on the PCB
CM Currents 30MHz E-field On A MPQ4430 Board
CM Currents 300MHz E-field On A MPQ4430 Board
CM Currents Inductor And SW Node

The node with the largest AC swing (switch node), the highest elevation (inductor) and largest area on the PCB dominates the RF E-Field emission.
CM Currents Inductor And SW Node

You can easily check this by placing a grounded E-Field shield over your assumed emitter.

![Diagram showing grounded E-Field shield over PCB]
CM Currents Inductor And SW Node

That E-Field shield can look more complicated

Monopole E-Field AV spread spectrum with top heat sink as shield
You can shield your inductor, but don’t forget the switch node areas on the PCB and MOSFET’s / ICs.

Best is to keep with a solid shield 1mm away from the inductor surface to reduce eddy current losses.
We learned that they are RF E-Field emitters.

That can be helped by electrical shielding.
RF Current Through The Inductor
Why is there so much more high frequency RF content than in the 90 degrees direction?
There is a high frequency (capacitive) conduction path right through the inductor.
RF Current Through The Inductor With Magnetic Probe

Since it is mainly capacitive (single digit pF) current will increase with frequency.

Because rectangular shaped waveform harmonics go down linear with frequency the resulting RF current is flat over frequency through the capacitive conduction path.

$$\text{SRF} = \frac{1}{2\pi \sqrt{CL}}$$
That’s the reason why the high frequency current comes up.
RF Current Through The Inductor With Magnetic Probe

To filter that we face several problems

We have a capacitive divider with Cout

Cout has Esl and Esr

There is magnetic coupling into Cout
You can use two capacitors in line with Vout in the middle to get partial cancellation of the magnetic coupling.

- Cinductor ~ 5pF
- Cout ~ 44µF
RF Current Through The Inductor With Magnetic Probe

Or you use a low profile inductor with low magnetic field and similar height to the $C_{out}$ capacitors
To reduce the magnetic coupling you could use a shielded inductor or (even partial) shield the inductor. That will help with the E-Field, too.
RF Current Through The Inductor With Magnetic Probe

If you have to go off the board with Vout (like often on LED supplies) you most likely will need to post filter.

For post filter everything that was said about input filters is valid.
How Resonances Transform Impedance

A resonance circuit has going through its resonance an impedance from zero to infinity so it will impedance match a certain point almost anything. If you have a narrow high Q EMI problem, look for the resonator which does the impedance transformantion.

A simple unconnected 57.5 MHz coil resonance circuit (68pF||112nH) creates this red peak.
How Resonances Transform Impedance

It’s the resonance with its phase shift and not the wire. With the capacitor removed there is nothing left.
What Is The EMI Bandwidth Of A Switch Node

The problem is that a switch mode PSU is a wideband noise source with its switching harmonics up to several hundreds MHz.
EMI Impact Of The Switching Frequency

If we double the switching frequency we double the number of switch transitions, so we double the EMI energy.

That is 3dB more (power is $10\times\log(2)$)

But with double the frequency you have now only half the bins to locate this energy.

As a result the spurs have 6dB more amplitude per doubling the frequency.
EMI Impact Of The Switching Frequency

- EMI energy is proportional to switching frequency.
  - 3db/octave = 10db/decade
  - With fixed frequency for every doubling, only half the bins are available for the energy result is 6dB/octave = 20db/decade

EMI Increase in dB vs. Increase in switching frequency

- EMI energy is proportional to switching frequency:
  - 3db/octave = 10db/decade
Design of Radiated Comb Generator Using Single-Ended Positive Emitter Coupled Logic (PECL) D Flip-Flop

Yoppy* and Muhammad I. Sudrajat

Comb Generator Spectrum
Comb Generator Spectrum
Comb Generator Pulse with a Real 20GHz scope from LeCroy

Pulse is about 400ps long with 1.2Vpp Slope faster than 200ps
Measured at center SMA connector terminated with the scope @ 50 Ohm
Comb Generator Spectrum 1MHz Pulse repetition

* Att: 15 dB
* Ref: 46.00 dBμV
* RBW: 120 kHz
* VBW: 10 MHz
* SWT: 700 ms

45 dBUV
Comb Generator Spectrum 500kHz Pulse repetition

- Att 15 dB
- Ref 42.00 dBµV
- RBW 120 kHz
- VBW 10 MHz
- SWT 700 ms

10MHz 65dBuV
1MHz 45dBuV
500kHz 39dBuV

39dBuV

Start 0.0 Hz  Stop 1.0 GHz
FSS Perfect Spread Spectrum Maximum Possible Attenuation

If the sweep would be perfect with infinitive speed the maximum attenuation, because we are talking power will be: 
\[ \text{Attenuation} = 10 \times \log_{10} \left( \frac{RBW}{fsw} \right) \]

<table>
<thead>
<tr>
<th>RBW [MHz]</th>
<th>0.15MHz-30MHz</th>
<th>30MHz-1GHz</th>
<th>&gt;1GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.009</td>
<td>0.12</td>
<td>1</td>
<td></td>
</tr>
<tr>
<td>fsw [MHz]</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>0.1</td>
<td>-10.5</td>
<td>0.0</td>
<td>0.0</td>
</tr>
<tr>
<td>0.2</td>
<td>-13.5</td>
<td>-2.2</td>
<td>0.0</td>
</tr>
<tr>
<td>0.4</td>
<td>-16.5</td>
<td>-5.2</td>
<td>0.0</td>
</tr>
<tr>
<td>1</td>
<td>-20.5</td>
<td>-9.2</td>
<td>0.0</td>
</tr>
<tr>
<td>2</td>
<td>-23.5</td>
<td>-12.2</td>
<td>-3.0</td>
</tr>
<tr>
<td>3</td>
<td>-25.2</td>
<td>-14.0</td>
<td>-4.8</td>
</tr>
<tr>
<td>5</td>
<td>-27.4</td>
<td>-16.2</td>
<td>-7.0</td>
</tr>
<tr>
<td>10</td>
<td>-30.5</td>
<td>-19.2</td>
<td>-10.0</td>
</tr>
<tr>
<td>100</td>
<td>-40.5</td>
<td>-29.2</td>
<td>-20.0</td>
</tr>
</tbody>
</table>

You can think how many bins with RBW size are empty between the harmonics where the spurs can hide
Probing A MPQ4430 With A Langer MFA-R 0.2-75 Magnetic Probe

![Image of MPQ4430 circuit board]

![Image of KL 3000 Efficiency Meter reading 12.3160V, 3.3367V, 0.6066A, 2.0620A, 0.591 W, Eff= 92.1%]

![Image of waveform measurement]

190
Probing a MPQ4430 with a Langer MFA-R 0.2-75 Magnetic Probe

12.3160V  3.3368V
0.6065A  2.0612A
0.592 W  Efficiency = 92.1%
Probing A MPQ4430 With A Langer MFA-R 0.2-75 Magnetic Probe

- Voltage: 12.3160V, 3.3369V
- Current: 0.5842A, 1.9892A
- Power: 0.558 W
- Efficiency: 92.2%
Probing A MPQ4430 With A Langer MFA-R 0.2-75 Magnetic Probe
Isolated Layout EMI And Hints

Automotive EMI Deep Dive Seminar

June, 2018
Isolated Supplies – E-field Dominated

Parts with high dV/dt voltage excitations like the switch node and all parts “moving” with it show E-Field radiation.

Now we are talking about this range
Isolated Supplies – E-field Dominated
Any “Plate” With High $dV/dt$ Creates An RF E-Field
Where Do The CM And DM Currents Flow?

L1
N
PE

Differential Mode
Common Mode

With kind approval © Lorandt Fökel @ Würth Elektronik eiSos®
The Common Mode Path

With kind approval © Lorandt Fökel @ Würth Elektronik eiSos®
If you are lucky and have a PE with complete encapsulation

Transformer coupling capacity

With kind approval © Lorandt Fölkel @ Würth Elektronik eiSos®
If you are unlucky and have an isolated unit, all common mode currents flow to a plate called the universe and get radiated.
Isolated Supplies – E-field Dominated

Transformer coupling capacity

You can cut that current path at any place

With kind approval © Lorandt Fökel @ Würth Elektronik eiSos®
Isolated Supplies – E-field Dominated

Transformer coupling capacity

Here you can use a common mode choke

At this place it is a bit more difficult
Isolated Supplies – E-field Dominated

Transformer coupling capacity

There must be a RF $dV/dt$ source to create this current

With kind approval © Lorandt Fökel @ Würth Elektronik eiSos®
Isolated Supplies – E-field Dominated

Transformer coupling capacity

Which capacitively couples into a lot of nodes
You cancel the common mode current, when you make the capacitive coupling symmetric.
You cancel the common mode current, when you make the capacitive coupling symmetric.
First make the parasitic of C1 to C4 as low as practical, then chose the CY cap at one or two of the C1 to C4 positions and optimize its values through common mode conducted emission measurement.
Isolated Supplies – E-field Dominated

That’s the reason off-line looks so “airy.” If E-Field is the dominant contributor, keep the board “open” / low capacitance to the universe.
Appendix M: Our New EMI Lab In Hangzhou

New MPS EMI test center in Hangzhou, China
Additional labs to be built in Detroit, Germany & Shanghai
DC/DC Buck Converter Review

Automotive EMI Deep Dive Seminar

June, 2018
Buck Converter Basics

Duty cycle (Q1) at SW-node:
\[ d.c. = \frac{V_{out}}{V_{in}} \Rightarrow T_{ON} = \frac{1}{f_{sw}} \times d.c. \]

\[ \Delta I_L = \frac{(V_{in} - V_{out})}{L} \times T_{ON} = V_{out}/L \times T_{Off} \]

Buck converter design rule:
\[ \Delta I_L \sim 30\% \text{ to } 40\% \text{ of } I_{out} \]

\[ L = \frac{(V_{in} - V_{out}) \times T_{on}}{0.4 \times I_{out}} \]
Buck Converter Basics

Inductor current waveform

\[ I_{\text{peak}} = I_{\text{out}} + \Delta I_L / 2 \rightarrow 1.2*I_{\text{out}} \]

\[ V_{\text{out ripple}}: \quad \Delta V_{\text{out}} \approx \Delta I_L \left( \text{ESR} + \frac{1}{(8*f_{\text{sw}} * C_0)} \right) \]

Input capacitor AC current:

\[ I_{\text{in}, RMS} \approx I_{\text{out}} \sqrt{\text{d.c.} \cdot (1 - \text{d.c.})} \]
Buck Converter Example

Example: 24V to 3.3V
2.5A Load with 500kHz

\[ \Delta I_L = 30\% \times 2.5A = 0.75A \]
\[ \rightarrow I_{\text{peak}} = 2.875A \]

\[ L = (3.3V + 0.1V) \times 1.71\mu s / 0.75A = 5.83V\mu s / 0.75A = 7.8\mu H \]
Buck Converter Example

Example: 24V to 3.3V
2.5A Load with 500kHz
MPQ4430 FC-QFN3x4

We use 10μH ⇒ XAL5050 41mΩ,
Isat=4.9A, IRMS=3.6A
I_peak = 2.5A + 0.3A = 2.8A (10μH →ΔI_L=0.6A)

Output ripple Target 10mVpp:
For this we need the Output Capacitor ESR< 10mV/0.6A=17mΩ
And the capacity Co > 0.6A/(8*0.5MHz*10mV) = 15μF
→ 22μF X7R ceramic output capacitor
Slope compensation

Loss estimation: at 2.5A load
$Iq \times V_{in} \sim 6.6mA \times 24V = 160mW$ (CCM-no load)
Top-FET: $(2.5A)^2 \times 0.1\Omega \times 0.15 = 94mW$
Bottom-FET: $(2.5A)^2 \times 0.05\Omega \times 0.85 = 266mW$
Transition losses: $\sim 100mW$ (guess)
Total: $620mW = 7.5\%$ of $P_{out}$
Temp rise: $0.62W \times 11^\circ C/W = 6.8^\circ C$ vs. Case

Target: IC Switch Current Limit has to be higher than 3A at 15\% Duty-Cycle

|SYNC input high voltage| $V_{SYNC\_HIGH}$| 1.8| 5.8| 7.3| A |
|Current limit| $I_{LIMIT\_HS}$| Duty cycle = 40\%| 4.7| 5.8| 7.3| A |
|Low-side valley current limit| $I_{LIMIT\_LS}$| $V_{OUT} = 3.3V, L = 4.7\mu H$| 3.1| 4.4| 5.0| |
|ZCD current| $I_{ZCD}$| 0.1| 0.1| 0.1| 0.1|

Min. Value vs. Temp. & distribution $1.1A$ lower

$MPQ4430$ Data Sheet Specifications
Real World: MPQ4430 Ev-board

3.5A 36V/40V Buck EMC Optimized Layout

- **Cout= 2x 22µF**
- XAL5050-10µH main coil and XAL4030-4.7µH in EMC Filter

Input EMC Filter and optional 555 for SSFM

- Two stage Input Filter

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EN GND SYNC GND PG

36V MAX UEMI

GND

VOUT

10nm

MPS
Monolithic Power Systems
MPQ4430 Demo Board
T-EVQ4430-L-01A
www.mps-power.com
Made in China
EMC Filter

R5 = 20Ω slows down rising SW Edge. Less ringing and lower high Frequency noise

<table>
<thead>
<tr>
<th>R4</th>
<th>Fsw</th>
</tr>
</thead>
<tbody>
<tr>
<td>249kΩ</td>
<td>350kHz</td>
</tr>
<tr>
<td>169kΩ</td>
<td>500kHz</td>
</tr>
<tr>
<td>82kΩ</td>
<td>1MHz</td>
</tr>
<tr>
<td>36.5kΩ</td>
<td>2MHz</td>
</tr>
<tr>
<td>27kΩ</td>
<td>2.5MHz</td>
</tr>
</tbody>
</table>
MPQ4430 EV-board Video

Vsw [5V/div]  Iind [500mA/div]
Real World Output Ripple

24V to 3.3V 2.5A with 22μF and 10μH inductor

CH1: V-Switch 5V/div
CH4: I_L 500mA/div
CH3 Vout 5mV/div-AC

ΔI_L→L=8.5µH!

ΔV=8mV

I_Out

Ch1 Freq 498.5kHz
Ch3 Pk–Pk 7.40mV
Ch4 Pk–Pk 761mA
Ch4 Max 2.87 A

Δ: 360mA @: 2.50 A

20 Oct 2017
18:00:36
24V To 3.3V With 2.5A Detailed Look

- \( L \) increases when \( I_L > I_{\text{out}} \), \( V_{\text{out}} \) rises.
- \( V_{\text{out}} \) has \( \sim 7 \text{mVpp} \) Ripple.
- \( V_{\text{out}} \) is effective capacity at 3.3V is about 50% of 44\( \mu \text{F} \)!!

- \( I_L < I_{\text{out}} \), \( V_{\text{out}} \) falls.

**Specifications:**
- \( I_{\text{out}} = 2.5 \text{A} \)
- SW
- Out AC
- \( I_L \): 1.7A Offset
- \( V_{\text{out}} \): ~7mVpp Ripple

**Details:**
- Ch1: Freq 498.6kHz
- Ch3: Pk-Pk 6.32mV
- Ch4: Pk-Pk 728mA

**Additional Notes:**
- \( \Delta I_L \rightarrow L = 8.5 \mu \text{H} \! 
- \Delta: 356mA
- @: 2.50 A

20 Oct 2017 18:04:39

MPS
Power Losses in Buck DC/DC:

- a) Switch $R_{ON} \times I^2$
- b) $I_{supp} \times V_{in}$
- c) Switch transition losses
- d) Coil losses

With $V_{in}$ from 12V to 30V, a) and d) will not change very much, but b) & c) does!

b) 12V (80mW, 1%); 24V (160mW, 2%), 30V (200mW, 2.4%)

High side FET transition losses:

$$P \sim f_{sw} \times V^2 \times I^2 \times C_{gd}$$

Changing $V_{in}$ from 12V to 24V will increase transition losses by 4x! At 2.6A we see a

$\Delta$Power loss of 190mW, 80mW of that comes from b), ~100mW are switching losses.
## MPQ4430 Test Results With XAL, XEL And XFL4020-2.2µh

<table>
<thead>
<tr>
<th>Type</th>
<th>XAL4020</th>
<th>XEL4020</th>
<th>XFL4020</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fsw</td>
<td>500kHz</td>
<td>1.5MHz</td>
<td>500kHz</td>
</tr>
<tr>
<td>Delta I_L [A]</td>
<td>2.76</td>
<td>1.16</td>
<td>2.66</td>
</tr>
<tr>
<td>Total Loss</td>
<td>1.246</td>
<td>1.86</td>
<td>1.225</td>
</tr>
</tbody>
</table>

- **XAL and XEL4020 show no saturation at 4A Ipk.**
- **With 500kHz XEL4020-2.2µH is best, Δ=20mW**
- **XFL4020-2.2µH at 500kHz**
  - Saturation! Increased losses!

**XFL4020 is ~70mW better @ 1.5MHz**

500kHz has around 600mW lower losses Than 1.5MHz
Same Test With MSS7331-3.0µH Ferrite Coil

<table>
<thead>
<tr>
<th>Type</th>
<th>MSS7331-3µH</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fsw</td>
<td>500kHz</td>
</tr>
<tr>
<td>Delta I_L [A]</td>
<td>2A sat 3.8Apk</td>
</tr>
<tr>
<td>Total Loss [W]</td>
<td>1.254</td>
</tr>
<tr>
<td>Delta T [°C]</td>
<td>26</td>
</tr>
</tbody>
</table>

Due to saturation and higher ripple current, coil gets hotter at 500kHz.
A Closer Look To SYNC Buck Power Stage

Is that true: “For better Efficiency, use lowest R_ON FETs!” ???

Check application duty cycle:
High Side (Q1) Conducts in ON time, for 24V to 3.3V ~ 14% of the time
If R_ON is decreased, Cg and Cgd increase

Low Side FET is conducting (1-d.c.) 86% of the time for the 24V to 3.3V circuit.
Q2 Miller capacitor Cgd is discharged by the coil!
### Some FET specifications:

<table>
<thead>
<tr>
<th>Drain-Source On-State Resistance</th>
<th>$R_{DS(on)}$</th>
<th>$V_{GS} = 10,V$, $I_D = 20,A$</th>
<th>$V_{GS} = 6,V$, $I_D = 15,A$</th>
<th>$V_{GS} = 4.5,V$, $I_D = 10,A$</th>
<th>$\Omega$</th>
<th>$R_8$ (k$\Omega$)</th>
<th>$R_9$ (k$\Omega$)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>0.0065</td>
<td>0.0080</td>
<td>0.0100</td>
<td></td>
<td>37.4 (1%)</td>
<td>12 (1%)</td>
</tr>
<tr>
<td>Total Gate Charge</td>
<td>$Q_g$</td>
<td>$V_{DS} = 30,V$, $V_{GS} = 10,V$, $I_D = 10,A$</td>
<td>20.8</td>
<td>32</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Gate-Source Charge</td>
<td>$Q_{gs}$</td>
<td>$V_{DS} = 30,V$, $V_{GS} = 6,V$, $I_D = 10,A$</td>
<td>12.1</td>
<td>18.5</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Gate-Drain Charge</td>
<td>$Q_{gd}$</td>
<td>$V_{DS} = 30,V$, $V_{GS} = 4.5,V$, $I_D = 10,A$</td>
<td>9.3</td>
<td>14</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Note: thermal PAD connects to PGND and AGND.
MPQ2908A Example: 12V 100W

Vin=32V

Vin=48V
Characterize PG-Pin Body Diode:

1) Measure $V_{\text{diode}}$ at two temperature points, 25°C and 100°C (with $V_{\text{in}}$ as in application)
2) Calculate temperature coefficient
3) Measure $V_{\text{diode}}$ at No Load
4) Measure $V_{\text{diode}}$ at 2.5A (500kHz, 24V)

MPQ4430 PG-Pin body diode has -1.55mV/K at 1mA current.
Real world circuit efficiency (warmed up > 15 min.) is:
\[ P_{out} = 3.326V \times 2.503A = 8.325W \]
\[ P_{in} = 24.008V \times 391.5mA = 9.400W \]
Efficiency is 88.56%.
Total \( P_d = 1.075W \); \( T_{amb} = 24^\circ C \); \( T_{PCB} \) rises to 33\(^\circ\)C; \( \Delta T_{PCB} \sim 10^\circ C \)
\( \Delta I_L = 0.84A \) Inductor is connected via short cables. Reduced heating of PCB.

If we assume the calculated Power Loss for the coil of 0.376W is correct, the IC has 0.7W losses.
With Theta J-C: 11°C/W \( \Rightarrow \Delta T = 7.7^\circ C \) this is above PCB temp. \( \Rightarrow T_J \sim 41^\circ C \)

with Theta J-A=48°C/W \( \Rightarrow \Delta T = 33.6^\circ C \) \( \Rightarrow T_J = 57.6^\circ C \)
What is correct??

\( \Delta V_d = 34mV \Rightarrow \Delta T_J = 22^\circ C \). \( \Rightarrow T_J = 46^\circ C \). This is reality!
Theta J-A for this set-up is 31.4°C/W
Theta J-C is \( (46^\circ C - 33.5^\circ C)/0.7W = 17.8^\circ C \) (problem to get exact PCB temp. at the IC)

Measurement on IC Top: \sim 40^\circ C \); Measurement at Coil: 43.3°C (\( \Delta T \sim 20^\circ C \))
Efficiency Of MPQ4430 With XAL4030 And XAL6030 3.3µH

24V to 3.3V with Fsw=1MHz

XAL6030 @3A 89.1%

XAL4030 @3A 88.2%

$\Delta P_{\text{loss}} = 106$ mW

1.814W

1.708W
Duty Cycle Range: Min./Max. Vin?

For a given $V_{out}$, the max. Vin in normal fixed frequency operation is:

Vin, max = $V_{out}/(d.c.,\text{min})$

The min Vin is:

Vin, min = $V_{out}/(d.c.,\text{max})$

Example MPQ4415 2.2MHz

<table>
<thead>
<tr>
<th>Maximum duty cycle</th>
<th>$D_{\text{MAX}}$</th>
<th>85</th>
<th>%</th>
</tr>
</thead>
<tbody>
<tr>
<td>Minimum on time ($^{(5)}$)</td>
<td>$T_{\text{ON,MIN}}$</td>
<td>46</td>
<td>ns</td>
</tr>
</tbody>
</table>

Vin, min: $3.3V/0.85 = 3.9V$

Vin, max: $3.3V/0.1 = 33V$

=10%

Example MPQ4425 1.5A LED driver with fixed $F_{\text{sw}}$ CCM

<table>
<thead>
<tr>
<th>Oscillator frequency</th>
<th>$f_{\text{SW}}$</th>
<th>$V_{FB}=100mV$</th>
<th>1800</th>
<th>2200</th>
<th>2600</th>
<th>kHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum duty cycle</td>
<td>$D_{\text{MAX}}$</td>
<td>$V_{FB}=100mV$</td>
<td>80</td>
<td>87</td>
<td>120</td>
<td>%</td>
</tr>
<tr>
<td>Minimum on time ($^{(5)}$)</td>
<td>$T_{\text{ON,MIN}}$</td>
<td>46</td>
<td>ns</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

$V_{\text{LED, max}} = \text{Vin} \times 0.8$ which is 9.6V for 12V input

$V_{\text{LED, min}} = \text{Vin} \times 0.1$ which is 1.2V for 12V input
MPQ4430 and similar parts with adj. fsw:

**Duty Cycle Range: Min./Max. Vin?**

But NO max. Duty Cycle Specification???

- Low Dropout Mode

---

**Minimum on time \( t_{ON,MIN} \):**

| Minimum on time (ns) | 80 ns |

---

**Output Voltage vs. Load Current**

Dropout Performance
(Set Nominal \( V_{OUT} > V_{IN} \))

- \( V_{IN}=5V \)
- \( V_{IN}=3.3V \)
For $I_{\text{Out}} < \Delta I_L/2$ buck enters discontinuous conduction mode.

MPQ4420 has special AAM mode for better LL efficiency.

AAM: $i_{pk}$ is clamped ~500mA -> $Fsw$ is reduced to keep output in regulation.
Short Circuit / Overload Test At 24V Input

Details:

Due to short output starts to drop. IC tries to keep output up with longer ON-cycle until CL is reached. When V-Fb < 660mV parts reduces Fsw to avoid inductor current runaway.

Short circuit entry from a 1A load condition. Frequency is fold back.

CH1: V-Switch 5V/div
CH4: I_L 1A/div
CH2: Vout 200mV/div-AC

CH2: Vout 2V/div
MPQ4420 Ev-board Video

Vsw [5V/div]  lind [1A/div]
Short Circuit Test At 14V – Small Coil

For lower current Buck converter with MPQ443x
Small inductor can be used

Examples: 14V to 3.3V 500mA
LQS 3015 -10µH $I_{\text{sat}}=0.87A; I_r=0.84A$ 230mΩ

3mm x 3mm x 1.5mm

<table>
<thead>
<tr>
<th>Properties</th>
<th>Test conditions</th>
<th>Value</th>
<th>Unit</th>
<th>Tol.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Inductance</td>
<td>100 kHz/1 V</td>
<td>L</td>
<td>µH</td>
<td>±20%</td>
</tr>
<tr>
<td>Rated current</td>
<td>$\Delta T = 40 K$</td>
<td>$I_R$</td>
<td>A</td>
<td>max.</td>
</tr>
<tr>
<td>Saturation current</td>
<td>$\Delta I/I &lt; 30%$</td>
<td>$I_{\text{sat}}$</td>
<td>A</td>
<td>typ.</td>
</tr>
<tr>
<td>DC Resistance</td>
<td>@ 20°C</td>
<td>$R_{\text{DC}}$</td>
<td>mΩ</td>
<td>±20%</td>
</tr>
<tr>
<td>Self resonant frequency</td>
<td></td>
<td>$f_{\text{res}}$</td>
<td>MHz</td>
<td>tvb.</td>
</tr>
</tbody>
</table>
Short Circuit Test At 14V – Small Coil

CH1: V-Switch 5V/div
CH4: I_L 2A/div
CH2: Vout 2V/div
CH3: Vin 10V/div

Δ: 4.70 A
Ω: 4.70 A

Ch1 Freq
423.7kHz

Ch1 Mean
657mV
Clipping negative

Ch3 Mean
14.0 V

CH1: V-Switch 5V/div
CH4: I_L 2A/div
CH2: Vout 2V/div
CH3: Vin 10V/div

2 Oct 2015
07:18:11
Short Circuit Test At 14V – Small Coil

Detail

CH1: V-Switch 5V/div
CH4: I_L 2A/div
CH2: Vout 2V/div
CH3: Vin 10V/div

Δ: 4.58 A
@: 4.58 A

Ch1 Freq
105.8kHz
Clipping
negative
Ch1 Mean
475mV
Clipping
negative
Ch3 Mean
14.5 V

2 Oct 2015
07:22:38
Short Circuit Test At 14V – Small Coil

Continuous Short

Hiccup Protection.

Ch1 Freq
82.96kHz
Low resolution
Ch1 Mean
371mV

Ch3 Mean
14.4 V

$I_{\text{average}} \approx 470\text{mA}$

In short
Ringing At Cin Due To Vin Hot-plug Via Long Cable

L, C resonant circuit with little damping! => High risk of Ringing!

Destructive high voltages 🌈 can build up!

Power source

Cable and EMC input filter coil(s)

Ri_source Few mΩ

L_cable

Several µH ~1µH/m cable

Rser_cable Few mΩ

ESR Few mΩ

Ceramic input capacitor at regulator

Few µF Less with Rising voltage!
Ringing At Cin Due To Vin Hot-plug Via Long Cable

Voltage at IC Vin On EV-Board

5V power supply Via 1m cable

6V Abs. max. IC was killed!

Ch1 Sp−Sp 8.08 V
Hot Plug 24V Via 2m Cable In 1µf 1206

Test set-up A:
- 24V
- 1µF 1206 50V
- 9.1kΩ

Probe

Test set-up B:
- 24V
- 1µF 1206 50V
- 9.1kΩ
- 22µF 35V
- Ø6.3mm; h=5.5mm

Probe

Al-Elco for damping

Tek Stop

A

30V ringing
Will destroy 40V IC

High Q and <<1µF effective Capacity cause 30V over-voltage

B

Safe for 36V/40V technology

With damping Al-Elco only 6V ringing

Reduced Q in resonance circuit
Real 22µF limit voltage change

Ch1 Max 52.8 V

Ch4 Max 10.4 A

15 Nov 2017 14:49:12

15 Nov 2017 16:56:09

MPS
Real World EVQ4572 HOT PLUG Via 2m Cable

Test with and without damping Al-Elco

Probe

CIN5 and CIN6: 470nF 0805
Hot Plug 24V Into EVQ4572 Via 2m Cable

Without Al-Elco:
36.8V → dangerous 13V overshoot

With 47µF Al-Elco:
25.6V → 1.6V overshoot
Current Mode Control

Error Amplifier Output ($V_c$) defines $I_{pk}$ for next cycle

If $V_c$ increases → $I_{pk}$ Increases
Duty Cycle increases…

Vout will NOT drop! 😊

This is of BIG advantage for fast input changes!!

- CM needs current sensing element!

Second – inner, fast current loop added
Gain ~ 300kΩ / (Rt + R1//R2) → increase gain
→ increases ft, but might reduce phase margin
Zero at fz = 1/(2π*Cff*R1) → increases Phase by 45° at fz
Example with Rt=24kΩ and no Cff!

Control Loop Bode Plot

Gain

Not good enough!

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Trace1</th>
<th>Trace2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cursor 1</td>
<td>67.484 kHz</td>
<td>-0.00 dB</td>
</tr>
<tr>
<td>Cursor 2</td>
<td>144.267 kHz</td>
<td>-8.871 dB</td>
</tr>
</tbody>
</table>

MPQ4423 24V To 3.3v 1.7Ω

Gain Control Loop Bode Plot

Phase Margin: 39°
Gain Margin: 8.8dB

Ft= 67kHz

Not good enough!
Example with $R_t=51\,\Omega$ and $C_{ff}=33\,pF$!

Phase curve is flat in the region of $f$ → robust against gain variations!

**Phase Margin:** $69^\circ$

**Gain Margin:** $15.1\,\text{dB}$

**Trace1**

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Trace1</th>
<th>Trace2</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cursor 1</td>
<td>37.496 kHz</td>
<td>-0.00 dB</td>
</tr>
<tr>
<td>Cursor 2</td>
<td>219.751 kHz</td>
<td>-15.146 dB</td>
</tr>
</tbody>
</table>
Slope Problem In Fixed Frequency Peak Current Mode

10V to 5V 500kHz 5µH
T_{ON}=1µS T_{OFF}=1µs ΔI=1A

Duty cycle=50%
Δi=0.1A
→Δt_{ON} = +0.1µs  →Δt_{OFF} = -0.1µs
Δi= -1A/µs*0.9µs= -0.9A
Δt_{ON}= +0.1µs
Δt_{OFF}= -0.1µs
Δi=0.2A→ 2x

Duty cycle=75%
Δi=0.1A
→Δt_{ON} = +0.2µs  →Δt_{OFF} = -0.2µs
Δi= -1.5A/µs*0.3µs= -0.45A
Δt_{ON}= +0.2µs
Δt_{OFF}= -0.2µs
Δi=0.4A→ 4x

At 90% duty cycle the effective gain is 10
Slope Problem In Fixed Frequency Peak Current Mode

Internal slope compensation correction curve takes care of this.

End user needs to be aware of duty-cycle dependant current limit

Current Limit vs. Duty Cycle

\[ I_{\text{LIMIT, HS (A)}} \]

\[ \text{DUTY CYCLE (\%)} \]

0 10 20 30 40 50 60 70 80 90 100
4.0 4.5 5.0 5.5 6.0 6.5
When $V_o$ is declined to $V_{ref}$, then high side MOS is turned on, the on time is constant which is decided by one shot section.

So the on-time duration is constant but off-time is variable.
COT Control

Merit of COT Control:

😊 Excellent load transient performance; ~ 4x faster than Fixed Frequency Current Mode
😊 Simple architecture---No need for compensation;
😊 Seamless transition between light load and heavy load;
😊 Do not need internal oscillator.

Demerit of COT Control:

楽しめる Need to generate slope on FB such as using Cout_ESR etc..
楽しめる The switching frequency is not constant due to variation off-time.
This is very difficult for design output filter and absolutely undesired in many sensitive systems.
When Vo decreases to Vref, then high side MOS is turned on, the on-time duration based on the input voltage, output voltage, so that it can achieve relative constant frequency operation. Adaptive COT control is derived from traditional COT control, using simple circuit in red line to make sure the switching frequency is constant.

\[ Q \cdot Ton \approx n \frac{Vo}{Vin} \]
\[ D = \frac{Ton}{Ts} \approx \frac{Vo}{Vin} \]
\[ \therefore Ts = \frac{1}{f_s} \approx \text{Constant} \]

The current source is proportion to Vin.
Adaptive COT Control

Merit of Adaptive COT Control:

- Excellent load transient performance;
- Simple architecture---No need for compensation;
- Seamless transition between light load and heavy load;
- Do not need internal oscillator.
- Fairly constant fs over Vin & Vo range.

Demerit of Adaptive COT Control:

- Need to generate slope on FB such as using Cout_ESR etc..

By using simple control circuit, adaptive COT control inherits all merits of traditional COT control and has constant switching frequency at the same time.
Detailed Implementation of Slope Comp. Circuit:

1) During HS-ON duration, a current \((V_{in}-V_{o})/R\) charges the cap; and a current \(V_{o}/R\) discharges the cap during HS-OFF duration. The ramp signal generated is in-phase with inductor current;

2) During load transient, the switching frequency changes, and it will generate a huge positive/negative ramp on C, which is still in-phase with \(I_{\text{inductor}}\), so the part is still stable.

Note: The impedance of C at the switching frequency should be several times smaller than the feedback resistors and make sure \(R>>R_1//R_2\) at the same time.
How does a EMI receiver or Spectrum analyzer work?

**Diagram:**
- **Mixer**
- **Resolution Bandwidth**
- **Peak detector**
- **Video Bandwidth**
- **Output**

**Flow:**
- **In** → **Mixer** → **Resolution Bandwidth** → **Peak detector** → **Video Bandwidth** → **Output**

**Sweeping Lo**
FFT analyzer operates in parallel after the mixer.
The resolution bandwidth filter is swept over the frequency range. That's a convolution operation if you remember signal class.
This is a convolution operation in the frequency domain.
Appendix A: EMI Measurement How Fast Can We Sweep?

The resolution bandwidth filter has limited response time in the time domain. Same as a low pass filter with the same bandwidth.

A 10Khz Gaussian filter has a Gaussian time response with 100us length.

So before moving to the next bin you need to wait over 100us.

There is still 4.3% amplitude at the Gaussian edges.

A typical time factor used is 3x the minimum time.

CISPR16-2-1 ask for 2x to 3x for near Gaussian filters.

That is the reason that the default video bandwidth setting on a spectrum analyzer is 3x the resolution bandwidth setting.
Appendix A: EMI Measurement How Fast Can We Sweep?

Let's look what CISPR16-2-1 says for sweep times.

<table>
<thead>
<tr>
<th>Frequency band</th>
<th>Scan time $T_s$ for peak detection</th>
<th>Scan time $T_s$ for quasi-peak detection</th>
</tr>
</thead>
<tbody>
<tr>
<td>A 9 kHz – 150 kHz</td>
<td>14.1 s</td>
<td>2820 s = 47 min</td>
</tr>
<tr>
<td>B 0.15 MHz – 30 MHz</td>
<td>2.985 s</td>
<td>5970 s = 99.5 min = 1 h 39 min</td>
</tr>
<tr>
<td>C/D 30 MHz – 1 000 MHz</td>
<td>0.97 s</td>
<td>19400 s = 323.3 min = 5 h 23 min</td>
</tr>
</tbody>
</table>

Band B uses a 9KHz resolution bandwidth
Band C/D uses 120KHz=
Using their own formula with \( k = 3 \) we get:

For the B band 1,106 seconds instead of 2,985
For the C/D band 0,202 seconds instead of 0,97
Appendix A: EMI Measurement How Fast Can We Sweep On Quasi Peak?

An additional filter is placed after the peak detector

<table>
<thead>
<tr>
<th></th>
<th>Rise time</th>
<th>Fall time</th>
<th>Meter time constant (now optional)</th>
</tr>
</thead>
<tbody>
<tr>
<td>For the B band</td>
<td>1ms</td>
<td>160ms</td>
<td>160ms</td>
</tr>
<tr>
<td>For the C/D band</td>
<td>1ms</td>
<td>550ms</td>
<td>100ms</td>
</tr>
</tbody>
</table>
Appendix A: EMI Measurement How Fast Can We Sweep On Quasi Peak?

<table>
<thead>
<tr>
<th>Rise time</th>
<th>Fall time</th>
<th>Meter time constant (now optional)</th>
</tr>
</thead>
<tbody>
<tr>
<td>For the B band</td>
<td>1ms</td>
<td>160ms</td>
</tr>
<tr>
<td>For the C/D band</td>
<td>1ms</td>
<td>550ms</td>
</tr>
</tbody>
</table>

With Rise time = R1*C1 and Fall time = R2*C1

If we forget the meter time constant the slowest response is the fall time. We need to wait 5x to 10x the time constant before we can move on.
Appendix A: EMI Measurement How Fast Can We Sweep On Quasi Peak?

Here is what a typical automotive OEM spec says

<table>
<thead>
<tr>
<th>BW (kHz)</th>
<th>PK</th>
<th>QP</th>
<th>AV</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Schrittweite Step size</td>
<td>Messzeit Measuring time</td>
<td>Schrittweite Step size</td>
</tr>
<tr>
<td></td>
<td>max.</td>
<td>min.</td>
<td>max.</td>
</tr>
<tr>
<td>9/10</td>
<td>≤ 0,5 x BW</td>
<td>50</td>
<td>≤ 5 x BW</td>
</tr>
<tr>
<td>120</td>
<td>≤ 0,5 x BW</td>
<td>5</td>
<td>≤ 5 x BW</td>
</tr>
<tr>
<td>1000</td>
<td>≤ 0,5 x BW</td>
<td>50</td>
<td>-</td>
</tr>
</tbody>
</table>
Appendix A: EMI Measurement How Fast Can We Sweep On Quasi Peak?

Let’s assume we wait 10x the time constant.

Band B: With a response of 1.6 seconds per 9KHz bin we can move through the 29.85MHz within 5307 seconds = 1h 28 min
CISPR16 states 1h 39 min

Which is close, because we do not know what assumptions they made.

Band C/D with a response of 5.5 seconds per 120kHz bin we can move through the 970MHz within 42626 seconds = 11h 51 min
CISPR16 calculates 5h 23 min.
They might be good with 5x time constant which gives only 40dB damping from the last bin result.
Appendix A: EMI Measurement How Fast Can We Sweep On Quasi Peak?

So Quasi peak slows down the possible sweep speed to a crawl with unrealistic long measurement times unless we use a FFT receiver which calculates the bins in parallel over its FFT bandwidth. Newer EMI receivers like Rohde & Schwarz series ESR, ESRP...

Since we know that peak measurement on a steady spectrum where the switching frequency is higher than the resolution bandwidth always is higher or equal than the Quasi peak measurement we can measure peak.
We know the time and frequency response of this first order low pass filter.
Appendix A: EMI Measurement How Fast Can We Sweep On Quasi Peak On A Non FFT Receiver?

A quasi peak detector slows down your frequency sweep in the C-band as if you had an 0.28 Hz Video bandwidth filter.
Appendix A: EMI Measurement Regime

We do a peak measurement first. If all peaks are below the quasi peak limit then the device has passed this test.

If some peak spurs are greater we scan with quasi peak filter setting around these spurs and check if we pass the quasi peak limits.

Same for average limits if they are different.

There is a catch. An automated test system will most likely pass a spread spectrum system since at some point the system can not differentiate the spread spectrum from noise.

However if you know it is spread spectrum CISPR 16 requires you to go slow.
Classic EMI receivers had needle instruments. To measure the annoyance factor for quasi peak and average time constants where defined in the old CISPR16-1-1 standard. Since there are discrepancies to modern instruments in the last revisions after 2010 the meter time constants were dropped.

**Table D.2 – Meter time constants and the corresponding video bandwidths and maximum scan rates**

<table>
<thead>
<tr>
<th></th>
<th>Band A</th>
<th>Band B</th>
<th>Bands C and D</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency range</td>
<td>9 kHz to 150 kHz</td>
<td>150 kHz to 30 MHz</td>
<td>30 MHz to 1 000 MHz</td>
</tr>
<tr>
<td>IF bandwidth $B_{\text{res}}$</td>
<td>200 Hz</td>
<td>9 kHz</td>
<td>120 kHz</td>
</tr>
<tr>
<td>Meter time constant</td>
<td>160 ms</td>
<td>160 ms</td>
<td>100 ms</td>
</tr>
<tr>
<td>Video bandwidth $B_{\text{video}}$</td>
<td>0,64 Hz</td>
<td>0,64 Hz</td>
<td>1 Hz</td>
</tr>
<tr>
<td>Maximum scan rate</td>
<td>8,9 s/kHz</td>
<td>172 s/MHz</td>
<td>8,3 s/MHz</td>
</tr>
</tbody>
</table>

Will give ultra long measurement times
Appendix A: EMI Measurement Now For Average?

If you have only a normal spectrum analyzer without special EMI average filter setting you can replace the average detector “memory” by decreasing the video bandwidth. To simulate the old dial receivers the video bandwidth setting would be in the 0.64Hz to 1Hz region which would slow down frequency scan to a crawl. But with typical switching regulators spectrums you dial down the video bandwidth only as far until you get a smooth readout.

See Appendix B CISPR16-2-1 for analyzer settings. In Appendix D CISPR16-2-1 you find examples for pulse suppression at 100 Hz video bandwidth.
Appendix A: EMI Measurement

General rule from signal theory: \( PK > QP > AV \)

If you don’t cheat and run too fast

You can always use the higher measuring Filter like PK for QK or AV limits

I use often combined limit lines with PK and QK limits and measure with PK wich gives the fastest sweep time.

For a fixed frequency DC-DC \( PK=QP=AV \) so it makes no difference.

For a spread spectrum DC-DC with reasonable modulation frequency and cyclic waveform like triangle @ 9kHz it is the same.

Only beware on digital pseudo random modulation. Their potential low frequency contents can make PK much higher than AV
Appendix A: EMI Measurement Fixed Frequency Switcher

A fixed frequency switch has a comb like spectrum which starts to die out above 300MHz.
Appendix A: EMI Measurement Fixed Frequency Switcher

For an ideal phase noise free fixed frequency switcher

\[ PK = QP = AV \]

Phase noise does rise with the harmonic number, so in reality >30 MHz the oscillator phase noise will widen the spurs and reduce its peaks.

Above 300-500MHz there is usual no measurement difference between a fixed frequency and spread spectrum switcher because oscillator phase noise of the >1000. harmonics make spread spectrum enough even if your part is fixed frequency.

That is not true for switchers synced to a high quality external oscillator like a crystal.
Appendix B: FSS Switcher With 100% AM Modulation 9khz RBW

So you will soon be in this region = 150KHz to 30MHz

\[ N \times \frac{f_{\text{mod}}}{N_0} \]

Appendix B: FSS Switcher With 100% AM Modulation 120kHz RBW

So you will soon be in this region

\[ \frac{N \times f_{\text{mod}}}{N_o} \]

What a low pass filter does from DC to its corner frequency does a bandpass filter for frequencies inside its passband.

So as long as the frequency stays inside the passband it is integrated up until saturation with $V_{out}(\text{filter}) = V_{in}(\text{filter})$. 
So if we sweep a carrier across a bandpass filter the signal will be integrated until saturation as long as it is within the passband.
The step response of a bandpass is \( \frac{1}{RBW} \) with \( RBW = \) the width \( F_H - F_L \) of the bandpass filter.

Impulse response of a 120KHz bandpass

If we want to get any attenuation at all we need to sweep our carrier faster through the RBW filter than its impulse response.
Appendix B: FSS RBW Band Pass Filter Behavior

We get no attenuation if the carrier stays longer than the bandpass impulse response time \( \frac{1}{RBW} \) inside the passband.
Appendix B: FSS RBW Band Pass Filter Behavior

The “speed” which the frequency changes need to be faster than

\[
\frac{RBW}{Impulse\ response\ time} = \frac{RBW}{RBW} = RBW^2 = a \text{ with be measured in } \frac{Hz}{s} \text{ or in } \frac{Periods}{s^2}
\]

According to CISPR16 / 22 / 25 for the bands to get any attenuation at all the frequency change speed or sweep need to be faster than

- 150kHz to 30MHz with RBW = 9kHz > 81MHz/s
- 30MHz to 1GHz with RBW = 120KHz > 14.4GHz/s
- > 1GHz with RBW = 1MHz > 1THz/s
Appendix C: Bench Side CISPR25 EMI Measurement Set Up

Goal:

Decent accurate set up for conducted and radiated emissions for DUT sizes of a demo board compared to a true CISPR25 chamber set up.

Small lab space, movable inside the lab and transportable with a passenger car or mail.

Affordable.
Appendix C: Bench Side CISPR25 EMI Measurement Set Up

Reliable CISPR25 measurements currently require a large absorber room (ALSE), costs a good 4 digit $ a day, costs windscreen time, can often not scheduled when needed and does not easy support add hock circuit modification.
Appendix C: Bench Side CISPR25 EMI Measurement Set Up

Automotive EMI is even more non-intuitive than the “normal” office equipment EMV.

The OEM and CISPR25 limits are very low compared to line powered CISPR22 style office equipment.

There are OEM limit areas which are less than a factor of 15 above the thermal noise of a 50 Ohm termination resistor. (i.e. 2 dBuV AV limit in VW81000 for the 2m band)

CISPR25 covers the influence of RF emitters to on board receivers and was developed from a military standard MIL-STD 461 (i.e. airplanes with receivers for navigation and comms).

The close proximity of emitters and receivers and the wide frequency range covers near-field and far-field effects.

Classical EMI measurements and theory covers far-field effects only, where radiation patterns, antenna factors etc. are well known and can be reproducible measured and simulated. CISPR22 (office equipment etc) uses official OATS (open area test site) with 10m between DUT and antenna for radiated emissions above 30MHz. For 10m distance anything above 5MHz can be regarded as far-field.

The automotive world with CISPR25 measures with 1m antenna distance to DUT or wire harness where frequencies below 50MHz are in the near-field area.
Appendix C: Near-field Vs. Far-field

At the emitter you have E-Field and H-Field with 90 degrees to each other. The field energy is reactive and bounces in and out of the emitting component.

In the far-field E and H fields are in phase. You have a traveling E and H field wave.

Appendix C: Near-field Vs. Far-field

Near-field and Far-field Fresnel zone

CISPR22 Receiver
CISPR25 Receiver

emitter

often in this area
Wavelength 150KHz to 30MHz is 2000m to 10m which is way above the 1m distance the Monopole rod antenna has to the setup. This is a complete capacitive coupling measurement between the 1m rod (few pF) and the DUT E-Field. This can be measured with a E-field antenna which can be closer than 1m to the DUT. The 1.5m cable is short compared to the wavelength and has not much influence.
Appendix C: CE (Conducted Emissions) Is Relative Simple

The low limit lines compared to ambient noise require a shielded environment.
Appendix C: CE Measurement

We need only a shielded box for the DUT and its 20cm cable. CISPR25 calls for the DUT 5cm above a conductive grounded table. So the bottom side is easy and will dominate the DUT / board and cable to GND capacitance. If we make enough space above the DUT that the capacities and with it the resonant frequencies are not too much lowered compared to an ALSE we can make CISPR25 standard conform CE tests.

The LISN is better shielded outside the box, since the RF measuring cable has to go to the external EMI receiver anyway. This way we save one RF feed through to the shielded enclosure.
Appendix C: Biconic Measurement 30mhz To 200mhz

Up to 200MHz the log per antenna measures the near-field radiation of the 1.5m micro strip line the supply cable forms which are mostly common mode currents. Those show up in the CE measurement too if you measure to those higher frequencies. Horizontal radiation is dominant.
Appendix C: Log Per Measurement >200mhz

Same set up of the log per antenna to biconic. Now the vertical component is the dominant radiation because the horizontal component is mostly confined close to the the 200-300 Ohm Micro strip line. The vertical radiation is mainly emitted direct form the DUT. Those still can be measured with a CE measurement since now the 200-400mm supply wires are a decent pick up for this high frequency range.
Appendix C: Bench Side CISPR25 EMI Measurement Set Up

We have a solid GND 5cm below the DUT and wiring which is easy to replicate. Capacitive effects below and above the DUT to the absorber room will determine resonance frequencies.

We can measure and correlate the Monopole E-field above the DUT with a decent distance where the E-field lines are almost flat. Assumed 10x10cm DUT board size 30-50cm height above it will do.

We can measure and correlate the conducted emissions, and radiated biconic and log per antenna measurements with a wideband (1GHz) CE LISN measurement.

In order not to change/reduce the capacitive resonance frequencies of the DUT, the E-Field lines should be close to the free space / large absorber room.

Under the DUT there is 5cm solid GND from the table.

Above for a 10x10cm DUT we should have 30-50cm free air space before any enclosure. On the side of the DUT it can be less since the E-Field lines are bent to the GND plane anyway.
Appendix C: Bench Side CISPR25 EMI Measurement Set Up

So a realistic enclosure would look like this:

Front view:

Actually E-field lines will always go perpendicular into a conductive enclosure but you get the idea.
Appendix C: Bench Side CISPR25 EMI Measurement Set Up

Side view:

- DUT
- E-field Antenna
- LISN
- Door
- EMI Receiver
- Absorber
- Loadresistor
- PSU

Diagram showing the setup for bench side CISPR25 EMI measurement.
Appendix C: Bench Side CISPR25 EMI Measurement Set Up

Side view:

- **DUT**
- **E-Field Antenna**
- **LISN**
- **EMI Receiver**
- **Absorber**
- **Loadbox**
The enclosure need to be RF sealed from all sides and conductive.

A watertight metal barrel is a good starting point. Steel has high losses over frequency and has excellent shielding capability.

Price of a new barrel is a small multiple of the material cost due to high automated production volume about 50 Euro / piece.

The problem is the lid connection and the door opening.

Size is ideal with \( d = 570 \text{mm} \) and \( h = 860 \text{mm} \)

The circumference seal ring will not be sufficient for the high frequency RF seal.
Appendix C: Bench Side CISPR25 EMI Measurement Set Up
Appendix C: Bench Side CISPR25 EMI Measurement Set Up

E-Field antenna (capacitive antenna) can be self build or bought from various sources. Google or look at your preferred auction platform for MiniWhip or bonito antenna. There are various designs from HAM Radio amateurs published. All have in common that they convert the high impedance of the antenna with a RF FET or MMIC. They all come with a bias-T with typ. 12V supply.

Replace with SMA connector

You increase the capacitive pick up with copper foil until you get similar results as with a Monopole in a CISPR25 ALSE

https://www.ebay.de/itm/MiniWhip-Active-Antenna-HF-LF-VLF-mini-whip-ham-radio-receiver-shortwave-swl-sdr/292558048041?hash=item441dd18b29:g:1i4AAOSw~FJZLor7
Appendix D: 5uh CISPR25 LISN For An Sealed Enclosure

Standard LISN are not good prepared to work in a small enclosure. They need to be of coaxial construction to get power and signals in and out a hermetic EMI sealed box. So I use a sealed enclosure and power goes in and out a coax PL259 connector (SO-239) which perfect fit a 4mm Banana to the inner conductor. Is good to >30A DC compared to a RF wise better N which is 5A only.

Here you screw it with a 19mm hole through the shielding of your enclosure. The 5-24 UNEF HEX nut is the same on a N-Connector
Appendix D: Coilcraft 6x SER2009-801MJB With Prot.
Appendix D: LISN For An Sealed Enclosure

So the most important factor for wideband operation besides high impedance inductors is the 50 Ohm path to the SMA connector which should have as low capacity as possible. That overrules even a 50 Ohm PCB design of this small board.

To protect the EMI receiver and other costly equipment it would be a bad idea to do away with the protection at the LISN output.
Appendix D: LISN For An Sealed Enclosure

According to CISPR25 the LISN is tested with 1Ohm shorting the VIN, 50Ohm at the RF_out and impedance tested at the DUT.
Appendix D: The Tolerance Is Given By CISPR16-4-1
Appendix D: A Look Into A Schwarzbeck LISN NNBM8124 Data Sheet

Abb. 4: Betrag der Impedanz an den Prüflingsklemmen (Kalibrieradapter KA 8125 erforderlich), BNC mit 50 Ω Abschluss, Speiseklemmen kurzgeschlossen

Fig. 4: Magnitude of impedance at EuT-Terminals (Calibration adapter KA 8125 required), BNC-Port terminated with 50 Ω, Mains terminals shorted
Appendix D: Compared To The Schwarzbeck LISN NNBM8124 Data Sheet We Are Well In Spec.
Appendix D: Isolation Is O.K.
Case is a Standard CTPE111/NEMA Aluminum die case made for the US market in 115x65x50mm.
Appendix D: LISN For An Sealed Enclosure

There are a few mounting holes to drill easiest on a mill with dials. Measurements always from the top opening since the die cast has a little angle inward.
Appendix D: LISN For An Sealed Enclosure

We need two SO 259 PTFE (2pf) connectors with 4x 3mm mounting holes. Do not even think of using central screw type SO 239 since they will turn and take the internal contacts with them... One hex nut 5/8x24UNCf and one long SMA connector PCB edge mount. 8x M3x12 screws 12x M3 nuts and two small PCB boards which holds the input filter and the output protection circuit. Coilcraft 6x SER2009-801 inductors soldered on the fly for low capacitance.
Appendix D: LISN For An Sealed Enclosure

2 Layer PCB for the protection circuit on the SO239 and the SMA out socket
Appendix D: LISN For An Sealed Enclosure

BOM:
The case there may be beside ebay.com some suppliers in USA
The UHF SO-239 / PL connector need to be with PTFE isolation for low capacitance (2pF range). Other dielectricums like Delrin POM etc have too high capacitance and low VSWR. The 19mm 5/8-24 UNEF hex nut is on UHF connector and N-type the same. Might be easiest to buy the cheapest connector with one or two screws and throw the connector away.
Appendix D: LISN For An Sealed Enclosure

Drill the case.
Mount both SO-239 connectors with M3 screws. Make sure the screw heads are not higher than the connector flange.
Mount the input PCB with the 4 MLCC and MOV and tighten the additional 4 M3 nuts.
Mount the output PCB. You might turn it a bit while fitting in.
Solder 20mm thick solid copper wire 2.5mm² into both SO-23Connect it to the middle filter input PCB
Solder a thin wire from the SO-239 connector to the output PCB connection to leave a bit room for connector movement. Otherwise the connection will break using the SO-239 connector.
Solder the 6x 820nH inductors together.
Solder them on the thick wire on the SO-239 connectors.
Screw the two 5mm screws and the 4x 4mm screws of the housing.
Test the input to output with a ohm meter. Test the SMA input that you see a diode drop from the 1N4148 in both directions. And that the SMA is isolated over its 100nF capacitor from the input and output.
1On one sample check with an VNA the RF performance. You need an PL to N adapter or whatever the VNA input has. A 50 Ohm SMA terminator and a 1 Ohm resistor shorting the input SO-239
Appendix E: Amplifier For Bench Side CISPR25 EMI Measurement Set Up

Problem is that most EMI receivers and especially spectrum analyzers have a relatively high noise figure. That means that their noise floor is much above the thermal noise a 50 Ohm resistor at room will produce at their input terminals. Spectrum analyzers give their noise in a DANL number which is recalculated to a 1Hz bandwidth in dBm. A 50 Ohm at room temperature has -174dBm so you can guessimate what to expect if DANL is given. However, the internal preamplifiers (if fitted) often have a limited dynamic range so you are better off using a state of the art external preamp anyway.

If should have a low single digit noise figure. And have a high dynamic range. This is measured by IP3. A bandwidth from 150kHz to 1GHz is usable.

There are lots of MMIC based LNA (low noise amplifiers) available. There is however one problem most share that they are relatively sensitive blowing up by large input transients. The reason is that often 5V abs. max. parts are used on 20V+ internal supplies with a larger drain/collector resistor were the self bias current sets the operating point to something <5V which is fine for the part. But now figure what happens when a large transient hits the input.

A preamp with its limited output swing is a god protection for the expensive analyzer mixer, so in any case a good choice.
Appendix E: Amplifier For Bench Side CISPR25 EMI Measurement Set Up

Spectrum analyzers have a limited dynamic range and are easily overloaded. If you have high signal energy even in a frequency band you don’t sweep through mixer and amplifier stages can saturate/rail and you get a lot of fake signals displayed. To check for that add passive attenuation (internal switch or external coax damping elements) and look that all signals are damped the same. Modern analyzers display sometimes overload but that is not bulletproof.

EMI receivers and spectrum analyzers made for regulatory compliance measurement have for this reason switched input filter banks to fight against out of band overloads.

However the good news for DC-DC is there is mainly a comb with switching frequency steps produced. With a decent EMI filter there should be no much higher spurs unless you create some high Q resonances.

So for DC-DC converter precompliance measurement you can get away with a standard spectrum analyzer if you use common engineering sense and a low noise figure preamp with good dynamic range.

I currently use a 2 stage LTC6433-15 design with good success.
Appendix F: Near-field Versus Far-field

The above picture shows, that any near-field source will turn into a far-field source at a distance of the wavelength $\lambda/2\pi$

So regardless if the radiation started as nearfield magnetic or nearfield electric field, at a distance of $\lambda/2\pi$ it is a perfect electromagnetic wave with the far-field impedance if the Universe of about $120\pi = 377$ Ohms. Those 377Ohms is the relationship of the E-field in V/m and the magnetic field in A/m
We have three different areas for a small antenna/wavelength structures as PSUs usually are. 

near-field or reactive region. Most energy stays local with the emitter and goes right back to it.

Transition Zone or Fresnel or intermediate region. Here waves start to travel

far-field or Fraunhofer region Waves are 377 Ohm planar and decay with 1/r
Appendix F: Near-field Versus Far-field

That's the most difficult zone for board EMI. Normally, you wouldn't expect any problem because the board is usually small against $\lambda/2\pi$.

Here anything will be reflected with a 90-degree phase shift to its source, and no energy escapes.
Appendix F: Near-field Versus Far-field

- 90 degree phase between E and H field
- <90 degree phase between E and H field
- 0 degree phase between E and H field
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