Investigation of Buck Converter Radiated Emissions (150 kHz – 30 MHz) Measured according to CISPR 25

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Investigation of Buck Converter Radiated Emissions (150 kHz – 30 MHz) Measured according to CISPR 25
To the love of my life!
SWITCHING POWER SUPPLIES

Generate noise at every frequency under the Sun (and some interstellar ones as well).

Every mode of noise transmission is present.

If you must use them you should filter, screen, keep them far away from sensitive circuits, and still worry!

James M. Bryant
Abstract

Electromagnetic compatibility and compliance with relevant standards is imperative for commercial success for any type of electronic equipment. Since more and more electronics are constantly added into today’s vehicles, this is a highly significant matter in the automotive business. The primary source of electric energy in an on-road vehicle is typically a 12 or 24 volt battery; this makes voltage step down converters ubiquitous in virtually any automotive electronic system. In strive for ever more environmental friendly and energy efficient solutions a switch mode power supply is most often the given choice when it comes to the task of voltage conversion. However, the use of switch mode power supplies presents a new set of challenges when it comes to successfully comply with the electromagnetic emission standards.

Knowledge and understanding about how different design parameters impact on EMC performance is key when few prototype runs and short time to market lies in focus. This text will investigate just how different layout design parameters affect the radiated emissions from a buck converter. Emphasis lies on radiated emissions in the lower frequency range up to a few MHz. Both computer simulations and practical measurements indicate the same thing; in the lower part of the frequency spectrum, when measured according to CISPR 25, radiated emissions from buck converters are dominated by voltage driven mechanisms. Along the way we will see how PCB layout alone can be responsible for differences in measured radiated emission levels of well over 20 dB.
Acknowledgements

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Takeshi Murase
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1 Introduction

In this chapter, as the title implies, I will briefly introduce a few topics which basically reflects what this text is all about. But before that follows a short section that outlines the contents of this text in a bit more detail. In this first section a few subject specific terms are used, which will be explained more in detail later throughout the text.

1.1 Scope

The purpose of this text is to investigate and try to gain an understanding of the mechanisms behind radiated emission measurements in the frequency range 150 kHz–30 MHz according to CIPSR 25. Even though the focus completely lies on the buck converter and the radiated emissions therefrom, the results are in many senses general and can be applied on a wide range of products. The primary goal is not to provide a design guide or a set of rules for an optimal buck converter design from the electromagnetic compatibility (EMC) point of view. The focus instead lies on understanding the emission source mechanisms in a buck converter and their contribution to radiated emissions, measured according to CISPR 25.

Of course, a solid understanding of these mechanisms provides the base knowledge for an EMC effective design. This approach means that the influences from physical structures are prioritized and the effect of different component choices and other circuit specific parameters are less, or not at all, addressed. We can think of it as a radiating system were we concentrate mainly on the antenna properties and care not so much about the amplifier that feeds the transmitting antenna.

Both several practical measurements and different computer simulations are performed and described. Although 30 MHz is the upper limit for the frequencies of interest in this investigation, often frequencies from 150 kHz up to a few MHz are of special interest, since it is in this frequency range we often find the switching frequency of a buck converter, and its first ten or so harmonics.
Since CISPR 25 is an automotive standard the voltage and power levels used in the simulations and measurements, mainly reflects those in a typical buck converter used in automotive applications.

But before we start with simulations and measurements some essential theory on electromagnetic interference (EMI) and buck converters will be discussed after this introductory chapter.

1.2 Short on EMC

The definition of electromagnetic compatibility according to the EMC directive 2004/108/EC [1] is as follows:

Electromagnetic compatibility means the ability of equipment to function satisfactorily in its electromagnetic environment without introducing intolerable electromagnetic disturbances to other equipment in that environment.

The purpose of the aforementioned EMC directive is to regulate the compatibility of equipment regarding EMC. The EMC directive itself states nothing about limits or measurement procedures, those issues are instead dealt with by several different commercial standards. If you are interested in more about standards and legislation regarding EMC, see for instance [2].

In EMC contexts the acronym EMI is also frequently encountered. EMI stands for electromagnetic interference and thus the two terms should not be carelessly interchanged. As a matter of fact you could state that EMC is the absence of EMI.

For an equipment to be electromagnetic compatible it not only has to have a certain amount of immunity against disturbances but also limited interference emission levels. The importance of EMC cannot be stated with enough emphasis. While not offering any obvious improvement or added functionality to a product, the lack of consideration to EMC issues during product design can be far more significant. The list of incidents related to EMI can be made long, and while sometimes even humorous the consequences can be all from simply annoying to indeed dangerous, all depending on the nature of the product or products involved.
1.3 Buck Converter Fundamentals

A buck converter is a switch mode DC to DC converter topology also known as voltage step-down converter or simply step down converter. As these latter names implies the output voltage of a buck converter is always less than or equal to the input voltage. The fact that the converter is of switch mode type means that the input voltage is “chopped” or abruptly switched on and off. This is done repeatedly with a frequency called the switching frequency. The switching frequency for a buck converter typically ranges from a couple of hundred kHz to a few MHz. It is this switching action that makes the buck converter a very energy efficient power converter. Unfortunately the switching has a disadvantage in the sense that it makes the converter a substantial noise source.

1.3.1 Basic Theory of Operation

Figure 1 shows the basic schematic of a buck converter.

As mentioned above, the switch is controlled in such a way that it is alternatively turned on and off, causing two distinct cycles. When the switch is closed, load current is supplied from the input source through the inductor, at the same time part of the energy supplied by the input source is stored as magnetic energy in the inductor $L$. Note that when the switch is closed the diode $D$ is reversed biased by the input voltage, and hence does not conduct any current. When the switch is open, no current flows from the input source. Instead load current is now supplied by the previous stored energy in the inductor; the diode is now forward biased and
thus provides a return path. The function of the capacitor $C_{out}$ is to smooth the output voltage.

If the current through the inductor never falls to zero, the converter is said to operate in continuous conduction mode, conversely if the inductor current falls to zero the converter is said to operate in discontinuous conduction mode. In the case of steady state and continuous conduction mode operation, the relation between input voltage, $V_{in}$ and output voltage, $V_{out}$ takes on a simple form and can be expressed as [3]:

$$V_{out} = D V_{in} \quad (1)$$

where $D$ denotes the duty cycle of the switching waveform. As $D$ varies between 0 and 1 we readily see that the output voltage cannot exceed the input voltage level.

A previously mentioned, the buck converter provides a means of efficient DC–DC-voltage conversion. To further improve efficiency, the diode $D$ in Figure 1 is sometimes replaced by a transistor. A buck converter with this type of active rectification is then recognized as a synchronous buck converter. In a practical converter the switch in Figure 1 is also realized with a transistor and a control circuitry monitors the actual output voltage and controls the transistor(s) to regulate the output voltage in a closed loop manner. For a more in-depth theory about switch-mode power supplies, [3] [4] [5], can all be worth a look.

1.4 Radiated Emissions & CISPR 25.

CISPR 25 [6] is a standard from the International Electrotechnical Commission. The title of the standard is:

Radio disturbance characteristics for the
protection of receivers used on board vehicles,
boats, and on devices –
Limits and methods of measurement

This standard establishes test methods for measuring the electromagnetic emissions from electrical components intended for use in vehicles. Limits for electromagnetic emissions are also set. According to CISPR 25 radiated emission measurements in the frequency range 150 kHz – 30 MHz shall be performed in an ALSE using a vertical monopole antenna. ALSE is an abbreviation for absorber lined shielded enclosure. Figure 2 shows such a measurement setup.
Figure 2. Example of test setup – rod antenna
1.4.1 Thoughts about the Measurement Setup

I find that the use of the measurement method suggested by CISPR 25 makes the interpretation of the results a bit tricky and is worth a second thought. The voltage from the monopole is measured with a spectrum analyzer, this voltage is then compensated for any cable loss and via the antenna factor converted to an electric field strength [V/m]. By the use of antenna factor we interpret the reading as if the antenna had been illuminated by a plane wave with its electric field (E-field) orientated along the monopole and electric field strength equal to the measured value.

A frequency range of 150 kHz – 30 MHz corresponds to a free space wavelength range of approximately 2 km down to 10 m. Therefore within this frequency range and with a measurement distance of only 1 m we are not even close to any valid far field approximations. The electromagnetic field in the vicinity of the measurement antenna is instead more likely to vary in both strength and direction; this fact may seem to make the measurement reading in [V/m] a little bit ambiguous. At first thought you might find it inconvenient that the standard calls for such a measurement method. But given the scope of CISPR 25 you realize that determine the effect on a structure that mimics the real situation is much more important than any equivalent plane wave field strength.
2 The Nature of EMI

In its simplest form, an EMI problem can be divided into three separate parts; a source of noise, a receiver of the noise and a coupling path for the noise between source and receiver. An important consequence of this approach is that, if one of the parts is removed the EMI problem cease to exist. The source of noise is often called the “aggressor” while the receiver of noise is then called the “victim”. Even though electronic products are EMC tested separately as a “victim” or an “aggressor”, i.e. susceptibility tests and emission tests, the problem with EMI is of course not limited to exist between different products, but can as well exist within a product. In the latter case the “victim” and “aggressor” are separate internal parts of the product.

2.1 Noise Coupling Mechanisms

In the world of EMC there is a distinction between radiated emissions and conducted emissions. Radiated emissions are measured using different types of antennas whereas conducted emissions are measured on the power cord of a product through a LISN. LISN stands for line impedance stabilization network and is a part of the measurement equipment and setup. A more detailed discussion about the LISN and its purpose can be found in [7].

Even though this text is focused entirely on radiated emissions, conducted noise is still an important issue due to the fact that a radiated noise problem is often composed of multiple mechanisms where conducted noise can be an intermediate coupling path or even the source itself. A common example of the latter is the case when conducted noise currents are allowed to flow out onto cables attached to a product, from where they later manifest themselves as radiated emissions.

We will now take a brief look at different physical mechanisms by which noise can couple in an electric circuit. Figure 3 illustrates the different modes of coupling.
2.1.1 Common Impedance Coupling

Common impedance coupling, sometimes also referred to as conductive coupling, occurs when two or more electrical circuits somewhere share a common impedance. Consider the case of Figure 4.

Suppose the objective is to monitor the voltage of $V_S$. As long as the switch is open we will read the actual voltage of $V_S$. But as soon as the switch is closed and
the resistance of $Z_C$ is not zero the reading will be corrupted by the current that
flows through $R_L$ and $Z_C$. Thus we see that the current in the circuit with $R_L$
through the common impedance $Z_C$ affects the result in the measuring circuit.
Clever grounding techniques can often prove valuable when dealing with common
impedance coupling issues.

2.1.2 Capacitive Coupling

This form of coupling can take place whenever there is a mutual capacitance
between parts of different circuits. Because it is the change of electric field that is
responsible for capacitive coupling, places in a circuit with high $du/dt$ should be
given special attention in this regard. To mitigate the effects of capacitive coupling
electric shielding or the use of so called Faraday screens are often employed.

2.1.3 Inductive Coupling

Inductive coupling can occur when there is a mutual inductance between different
electrical circuits. The phenomenon of inductive coupling relies on the fact that
there is a changing magnetic field. Because electric currents are closely related to
magnetic fields, current loops with high $di/dt$ should be given extra consideration
with respect to inductive coupling. Magnetic shielding can be difficult, especially at
lower frequencies. Instead often other approaches, such as careful placement and
orientation of inductive components, are used in order to minimize problems with
inductive coupling.

2.1.4 Electromagnetic Radiation

Finally, noise from a circuit can also propagate as electromagnetic waves and then
be picked up by another electric circuit. Even though neither capacitive coupling
nor inductive coupling relies on a galvanic connection between the circuits
involved, strictly speaking they are not radiated emissions but may instead be
referred to as near field coupling or induced coupling. In EMC standards and
regulations, no such distinction is made and they are all lumped under the term
radiated emissions. As a consequence, in this text I use the term radiated emissions in a
more loose sense and the words shall therefore not always be interpreted as “true”
radiated emissions. This fact can sometimes, not only in this text, cause a little bit
of confusion.

Compared to near field coupling, electromagnetic radiation can reach over far
longer distances. Electromagnetic radiation needs “antennas” in order to be
transmitted and received, therefore much of the countermeasures to tackle radiated noise problems are about minimizing the efficiency and excitation of unintended antennas in a system. For an in-depth theory on antennas see for instance [8] or [9].

2.2 Putting Things into Perspective

When talking about radiated emissions; electric field strengths are often given in [dB(µV/m)]. As an example, the CISPR 25 class 5 peak limit in the frequency range; 30 MHz – 54 MHz is 36 dB(µV/m), or approximately 63 µV/m. The quasi-peak level limit is 23 dB(µV/m), or just slightly over 14 µV/m. For a discussion about peak, quasi-peak and average values, see for instance [2].

To gain some insight into these levels let us do some simple calculations. For an electrically short dipole, or Hertzian dipole in free space, the maximum electric far field strength can be expressed as [7]:

\[ |E_{\text{max}}| = \frac{\eta_0 \beta_0 |\hat{I}| d l}{4\pi r} \]  

(2)

where

- \( \eta_0 \) = Intrinsic impedance of free space \( \approx 120\pi \) [Ω]
- \( \beta_0 \) = Free space wavenumber \( = 2\pi/\lambda_0 \) [m⁻¹]
- \( \lambda_0 \) = Free space wavelength [m]
- \( \hat{I} \) = Current amplitude [A]
- \( d l \) = Length of dipole [m]
- \( r \) = Distance [m]

Using (2) we find that if the dipole is 1 m long, a field strength of 36 dB(µV/m) at a distance of 1 m (consistent with the CISPR 25 measurement setup) corresponds to a current of 3.35 µA at 30 MHz. The corresponding number for the quasi-peak limit level of 23 dB(µV/m) is only 0.75 µA. Equation (2) is derived under the assumption that the current distribution along the dipole is constant.
A constant current distribution implicates a non-zero current at the endpoints of the dipole, which may not be very realistic in a practical case\(^1\), if instead a triangular current distribution is assumed; the maximum electric field strength is obtained by simply dividing the expression in (2) by a factor of two. An intuitive explanation for this can be acquired by observing that the average value along the dipole with triangular current distribution is one half of the peak current. See Figure 5.

![Figure 5. Triangular current distribution along dipole](image)

It is also important to remember that the expression in (2) is valid for far field only. Even though there is no sharp distinction between near and far field, the boundary in the case of the field from a Hertzian dipole is often taken to be \(\lambda_0/6\), where \(\lambda_0\) is the wavelength, which in turn is related to frequency \(f_0\) as:

\[
\lambda_0 = \frac{c_0}{f_0} \quad (3)
\]

where

\[c_0 = \text{speed of light in free space} \approx 3 \cdot 10^8 \text{ m/s}.
\]

Thus a frequency of 30 MHz yields a free space wavelength of 10 m. As stated earlier a Hertzian dipole is electrically short. Electrical length is a term often encountered in electromagnetic contexts; this length is expressed in wavelengths

\(^1\) By capacitively end-loading a dipole antenna the current at the outer parts of the dipole can be increased.
which in turn depend on propagation velocity and the frequency of the electromagnetic wave. Furthermore a physical structure can be considered electrically small or short when its size or length is just a tiny fraction of a wavelength. Likewise, the term electrically long implies a physical length approaching the order of a wavelength.

Despite the fact that the model used for the above field strength calculations is not applicable in a practical case or even theoretically (a distance of 1 m may not be enough for (2) to be accurate enough at 30 MHz), we can still make the following valuable conclusion: if we have an electrical equipment with attached cables and somehow drive a common mode current of only a few microamperes onto those cables, that alone can be enough to violate EMC compliance of the product. This is the reason why common mode currents are almost always so much more significant than differential mode currents when it comes to radiated emissions, see [10].

2.2.1 Non Ideal Components & Parasitics

Many EMC issues cannot be resolved or explained by only looking at the electric schematic of a product. Or put another way; two different units built from the exact same schematic can possess vastly different EMC characteristics. The only explanation to this is that properties that are not seen in the schematic are in fact affecting EMC. Because of this fact, you can sometimes hear things like “the hidden schematic” or even the word “magic” when EMC or EMI is discussed. Of course there is no magic involved in EMC, although it can surely seem like that from time to time. But the phrase “the hidden schematic” quite accurately describes what EMC issues is often all about.

As many radiated emission problems are related to high frequencies, and the fact that even small currents can be enough to cause trouble, the non-ideal behavior of components and parasitic elements are no longer negligible, regardless of the fact that they may not affect the functionality of the device itself. Perhaps the single most important component when used, which affects EMC is the printed circuit board, or PCB. Because the PCB is unique for every product and a lot of EMC properties depend on the PCB, the PCB design is a step where there are lots of possibilities to affect and hopefully improve EMC performance. I have seen firsthand how a bad layout can ruin an otherwise working design. Due to this significant role of the PCB, the impact on radiated emission levels of different
PCB designs were experimentally investigated, and the results will be presented later in this text.

### 2.2.2 The Importance of Dominance

Since electromagnetic noise can originate from several different sources and couple by multiple mechanisms, an EMI problem is typically composed of noise with many contributing factors. Moreover, as the individual contributions to the total emitted noise can be very uneven, understanding the dominant effect is a key factor when seeking for effective measures [11]. To explore this concept, a simple oscillating circuit according to the schematic in Figure 6 was built [12].

![Figure 6. 555 astable multivibrator schematic](image)

This circuit produces a square wave at the output pin of $U1$ which in turn drives the gate of $Q1$ to produce a pulsating current through resistor $R3$. The physical implementation of the circuit can be seen in Figure 7.
The whip in Figure 7 is 45 mm long and connected to the drain of Q1. The pulsating current loop can be extended by the loop in Figure 7, which is circular with a diameter of 45 mm, the loop extension is made at the upper end of R3 in Figure 6. When the current loop was not extended, the loop was simply removed and the circuit closed at the bottom of the loop. The radiated emission from this circuit with different configurations was then measured in a shielded enclosure using an active monopole. A more detailed description of the measurement setup is to be found in appendix A. For all measurements described in this section the circuit in Figure 7 was placed 1.2 m from the active monopole and fed by 1 m long cables from the LISNs. A 10 dB attenuator was also connected at the input of the spectrum analyzer. Table 1 shows the different parameters for the first set of measurements.

<table>
<thead>
<tr>
<th>#</th>
<th>R3 [Ω]</th>
<th>V_in [V]</th>
<th>Loop</th>
<th>Whip</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>330</td>
<td>12.4</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td>2</td>
<td>330</td>
<td>5.7</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td>3</td>
<td>330</td>
<td>12.4</td>
<td>Yes</td>
<td>No</td>
</tr>
<tr>
<td>4</td>
<td>330</td>
<td>12.4</td>
<td>No</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Table 1. Parameters for measurements in Figure 8

A “Yes” in the column “Loop” means that the pulsating current loop was extended by the loop in Figure 7 and the column “Whip” indicates weather or not
the *whip* in Figure 7 was mounted. A plot of the average value measured by the spectrum analyzer can be seen in Figure 8.

![Figure 8. Radiated emission measurement corresponding to Table 1](image)

From this figure we see that the difference in level between $T\# [1]$ and $T\# [2]$ correlates very well to the difference in supply voltage, at least for the first three or so peaks, which are the fundamental and the two first harmonics. The frequency shift in $T\# [2]$ is due to the fact that the oscillator frequency in this simple oscillator is somewhat affected by the supply voltage level. But since in this case we are only interested in relative differences the frequency shift is of little concern.

Next we notice that $T\# [3]$ which corresponds to the circuit with extended loop area shows no increase in radiated emission levels but rather a small decrease$^2$. This suggests that radiation from the current loop is not the dominant source in this case. On the other hand looking at $T\# [4]$ we see that the addition of the *whip* quite substantially increased the radiated emissions. These facts together with the observation that the radiated emissions seem to be proportional to supply voltage indicates that it is indeed the voltage that is the dominating parameter in this case.

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$^2$ The reason to this decrease may be understood by studying section 4.5.1 later in this text.
To further investigate the impact on radiated emissions by different parameters another set of measurements according to Table 2 was done.

<table>
<thead>
<tr>
<th>#</th>
<th>R3 [Ω]</th>
<th>V(_{\text{in}}) [V]</th>
<th>Loop</th>
<th>Whip</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>68</td>
<td>12.4</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td>2</td>
<td>27</td>
<td>12.4</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td>3</td>
<td>68</td>
<td>12.4</td>
<td>Yes</td>
<td>No</td>
</tr>
<tr>
<td>4</td>
<td>68</td>
<td>12.4</td>
<td>No</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Table 2. Parameters for measurements in Figure 9

This time the supply voltage was held constant for all the measurements and the pulsating current amplitude was changed by changing the resistor R3 in Figure 6. The result of these measurements can be seen in Figure 9.

Figure 9. Radiated emission measurement corresponding to Table 2

Again looking only at the first peaks we see that \(T^{\#}[1]\) and \(T^{\#}[2]\) shows no big difference between each other regardless of that the resistance is more than halved, and the current therefore more than doubled for \(T^{\#}[2]\) compared to \(T^{\#}[1]\).

Also looking back at Figure 8 we notice no increase in radiated emission levels corresponding to the resistance (and consequently current) ratio; which is more than 21 dB between \(T^{\#}[1]\) in Figure 8 and \(T^{\#}[2]\) in Figure 9. This fact reinforces
the impression that radiation emanating from the current loop is not the dominant source in this case.

Once again we see the same effect of adding the *loop* and *whip* in T# [3] and T# [4] respectively. Thus in these cases the contribution from the current loop seems to be so weak that an increase in loop area, except not contributing to an increase in emission levels even make the emission levels lower due to other more dominant mechanisms related to the physical change of the circuit.

We have only looked at the first peaks in the emission spectra, and a closer look at the higher harmonics reveals that the relative difference between different traces may not be that consistent. To understand why we have concentrated only on the first peaks, and to explain the discrepancy for the higher harmonics we turn to Fourier analysis. A rectangular pulse train with amplitude ±0.5 and duty cycle, \(D\) can be composed of a series of sinusoids with amplitudes \(a_n\) according to [13]:

\[ a_n = \frac{2}{n\pi} \sin(n\pi D) \]  

(4)

where \(n\) indicates the multiplication factor of the fundamental frequency. Using equation (4) the absolute value of the 15 first amplitudes, for two waveforms with slightly different duty cycle were calculated. The result is presented in Figure 10.

![Figure 10. Pulse train Fourier series coefficients](image)

Here we see that even though the duty cycle difference is quite small the relative difference between the higher harmonics can be large. Therefore a minor change
in duty cycle can manifest itself as a substantial change in radiated emission levels for the higher harmonics. Due to this fact and since the circuit in Figure 6 is quite sensitive to changes in different parameters; we relied on and analyzed the behavior of the first few peaks only.
3 Buck Converters & Radiated Emissions

The fundamental physical mechanism behind electromagnetic radiation is the acceleration (or deceleration) of electric charge. Since electric current involves movement of electric charge, all alternating currents have the potential to radiate, and so they will. In this chapter we will take a closer look at the possible sources of radiation from a buck converter. Some commonly used techniques to mitigate this unwanted radiation will also be looked at.

3.1 Common Mode & Differential Mode Currents

As pointed out in section 2.2; common mode currents are of major concern when dealing with radiated emissions. To clarify the difference between common mode currents and differential mode currents let us take a look at the simple lumped circuit model in Figure 11.

![Figure 11. Illustration of common mode and differential mode currents](image)
With the currents defined as in Figure 11 we obtain the following equations:

\[ I_1 = I_d + I_c \quad (5) \]
\[ I_2 = I_d - I_c \quad (6) \]

Combining equations (5) and (6) yields:

\[ I_d = \frac{I_1 + I_2}{2} \quad (7) \]
\[ I_c = \frac{I_1 - I_2}{2} \quad (8) \]

From equation (8) we see that if \( I_1 \) equals \( I_2 \) the common mode current is zero, then from equation (7) we also see that \( I_d \) is equal to \( I_1 \) (which is equal to \( I_2 \)) and we say that the circuit is fully balanced. The differential mode current \( I_d \) is usually the only current considered when a circuit is designed and is also a functional or wanted current.

In order to understand how unwanted common mode currents can occur, we again turn our attention to Figure 11. If the parasitic capacitance \( C_{p1} = C_{p2} = C_{p3} = C_{p4} \) the potentials at the two ground symbol nodes will be exactly the same and no common mode current will flow. As soon as the ratio between \( C_{p1} \) and \( C_{p2} \) is not equal to the ratio between \( C_{p3} \) and \( C_{p4} \), this asymmetry will give rise to a common mode current. I want to point out that the situation in Figure 11 is just one example of how common mode currents may be created, but in general asymmetries of one or another kind are involved. As practical implemented circuits as good as always exhibit asymmetric properties the possibility of common mode current generation is also present.

### 3.2 Sources of Radiation

Several instructive papers related to electromagnetic radiation from PCBs have been presented, [14] [15] [16], just to refer to a few. Those papers are mostly concerned with radiation originating from common mode currents on cables attached to a PCB. Two types of mechanisms driving these common mode currents are identified; namely voltage driven sources and current driven sources.
3.2.1 Current Driven Source

To understand the current driven source mechanism we take a look at Figure 12. The schematic is a model of a PCB with attached cables; \( L_{\text{trace}} \) represents the inductance associated with the PCB signal trace which interconnects the signal source \( V_s \) and load \( R_L \), and \( L_{\text{return}} \) represents the inductance associated with the return path. The signal return path is often through a local reference plane. Even though often small, the inductance associated with a finite sized reference plane is not zero [17].

![Figure 12. Illustration of the current driven source mechanism](image)

When a time varying current is passed through the load, this non-zero inductance associated with the return path will give rise to a voltage, which in turn can drive the attached cables as the two arms of a dipole. \( C_{\text{ant}} \) symbolizes the input capacitance of this unintended antenna.

Despite the fact that you quite often hear things like “the inductance of a wire” or “the inductance of a plane”, inductance is a concept with respect to a complete current loop. What is often probably meant by those types of statements is internal inductance or more likely partial inductance, which by the way is not the same\(^3\). To avoid confusion I used the word “associated” in conjunction with the inductances in Figure 12 to emphasize that they are not loop-inductances, but rather based on the concept of partial inductance [18].

---

\(^3\) The internal inductance of a wire with circular cross section and uniform current distribution is independent of the wire radius, whereas the partial inductance of the same wire is not.
3.2.2 Voltage Driven Source

The mechanism behind a voltage driven source can be explained with the help of Figure 13. In this figure $C_{ant}$ represents the capacitance between the upper signal trace and the attached cable. When the upper signal trace exhibits a time varying voltage, a common mode current can be driven onto the attached cable. As opposed to the case with a current driven source, the amount of common mode current driven onto the attached cable is not directly dependent on the load current, but rather on the voltage on the signal trace.

![Figure 13. Illustration of the voltage driven source mechanism](image)

Depending on load and source configuration, either the current driven or voltage driven source mechanism can be predominant.

3.2.3 A Comparison between the Two Source Mechanisms

For frequencies below the resonance of the unintentional antennas made up of the PCBs and attached wires, and where the structures are still electrically small, we can approximate the antenna input impedances as mainly capacitive, as depicted in Figure 12 and Figure 13. This means that the common mode current can be assumed proportional to the frequency of the common mode driving voltage. If, in the current driven case, the inductive reactance $(j\omega(L_{trace} + L_{return}))$ is much less than the load resistance $R_L$, the load current is mainly determined by the source voltage $V_s$ and the load resistance $R_L$, and thus can be approximated as frequency independent.
Moreover, since the driving voltage (voltage over $I_{\text{trace}}$) in the current driven case is proportional to the frequency of the load current, the common mode current driven onto the cables becomes proportional to the signal frequency squared, or will vary as 40 dB/decade with frequency. Now, in the voltage driven case, assuming that the driving voltage (signal trace voltage) is independent of load current, the common mode current driven onto the cable is proportional to signal frequency, or will vary as 20 dB/decade with frequency. This different frequency characteristic between the two source mechanisms may be utilized when trying to diagnose an EMI problem.

### 3.2.4 Loop Antennas

In addition to act as the driving source for a common mode current, as described in section 3.2.1, any time varying current loop will radiate itself. The efficiency of a loop antenna will greatly depend on its electrical size. For an electrically small circular loop in free space the maximum electric far field strength can be expressed as [8]:

$$
|E_{\text{max}}| = \frac{\beta_0 A \omega \mu_0 \dot{I}}{4 \pi r} \quad (9)
$$

where

$\beta_0$ = Free space wavenumber $= 2\pi/\lambda_0$ [m$^{-1}$]

$\lambda_0$ = Free space wavelength [m]

$\dot{I}$ = Current amplitude [A]

$A$ = Area of loop [m$^2$]

$\omega$ = Angular frequency [s$^{-1}$]

$\mu_0$ = Permeability of free space [H/m]

$r$ = Distance [m]

For a pair of electrically short wires with length $L$ and separation $d$, carrying a differential current $\dot{I}$, equation (9) can still be used to approximate the maximum electric far field strength. The area $A$ of the loop is then simply the length $L$ times the separation $d$. 

In the case of two 1 m long wires spaced 1 mm apart; using this equation we find that a current of 5.33 mA with a frequency of 30 MHz corresponds to an electric field strength of 36 dB(µV/m) at a distance of 1 m. Remember that the corresponding common mode current on a 1 m long wire, as calculated in section 2.2 was only 3.35 µA. This is a current magnitude ratio of more than three decades, or some 64 dB, for the same electric field strength. Yet again we see the significance of common mode currents when it comes to radiated emissions.

### 3.3 Sources of Radiation in a Buck Converter

We will now take a closer look at the currents and voltages in a buck converter and examine the possible sources of radiation. Figure 14 shows the same basic schematic as in Figure 1, but with added markings to indicate the different current paths and also the voltage switch node.

![Figure 14. Currents and voltages in a buck converter](image)

As explained in section 1.3.1, current will flow in one of two loops\(^4\), depending on whether the switch is closed or not. These two loops are marked in Figure 14. Part of these loops overlap each other and will carry a triangular shaped current, called inductor ripple current\(^5\). The path difference between the two loops makes a third loop which is indicated with a dashed line, and it is in this loop we will find current with the highest \(di/dt\).

---

\(^4\) An additional current loop exists between the output capacitor and the load.

\(^5\) This triangular shaped current is superimposed on a DC load current.
See Figure 15 for a typical ideal waveform of this current.

![Figure 15. Typical input current waveform](image)

This switching current may be the cause of electromagnetic radiation, both as the driving source for common mode currents and by direct loop radiation, as described in section 3.2.1 and 3.2.4 respectively.

The node encircled in red in Figure 14, connecting one side of the switch, the cathode of the diode and one end of the inductor is called the switch node. When the switch is closed, the potential at this node equals the potential at the input node. On the other hand, when the switch is open, the potential at the switch node is one diode forward voltage drop below the input and output common reference node. This means that the potential at the switch node will ideally be a square wave with a voltage amplitude approximately equal to the input voltage. The high $\frac{du}{dt}$ at the switch node makes it another potential source of radiated emissions.

In a practical converter, the existence of parasitic elements and non-idealities of components will add over and undershoot to this square waveform, which can then cause additional radiated emissions, most of this additional emissions will occur much higher up in the frequency spectrum compared to the fundamental frequency of the square wave [19] [20]. These over and undershoots on the square wave are commonly referred to as ringing.

### 3.4 Common EMI Mitigation Techniques

Depending on at which stage in a design process the question of EMC is addressed, or an EMI problem is discovered, the designer has more or less freedom to affect the design and consequently a better or worse chance of successfully and effectively prevent or solve a problem.
Suppressing the noise source itself or contain the noise to a limited region close to
the source is generally the preferred option, if possible. Once noise has coupled
into different parts of a system, controlling the emissions become much of a bigger
issue. An EMI problem discovered late in a design process may force the designer
to use last resort band-aid solutions; such as extensive filtering and or shielding.
Actions that can be very cost ineffective and perhaps unnecessary if EMC had
been considered in an earlier stage of the design.

Now follows a short survey of some commonly used techniques for controlling
EMI. PCB layout, being a crucial factor for controlling EMI is left out here but will
be addressed in more detail later throughout this text.

### 3.4.1 Rise & Fall Time Control

Rise and fall time control is entirely about to limit the excitation of harmonic
content in a signal, i.e. source-suppression. Square waves or digital signals are
present not only in buck converters, but in the vast majority of electronic circuits
designed today. Square waves have a broad frequency spectrum that stems from
the sharp edges in the waveform. A practical square wave can be approximated as a
signal with trapezoidal waveform. With the aid of Fourier analysis an upper
boundary for the spectral content of a trapezoidal waveform with equal rise and
fall times can be obtained [7]. The result can be summarized as seen in Figure 16.

![Figure 16. Trapezoidal waveform spectral envelope](image-url)
The notations in Figure 16 are:

\( A = \) Waveform amplitude

\( T = \) Period time

\( t = \) Rise and fall time

\( \tau = \) Pulse width at the 50 \% points

From this figure we see that slowing down the rise and fall time will push the break point \( f_2 \) to a lower frequency and thus lower the upper boundary for the harmonic content above this frequency.

In a buck converter, slowing down rise and fall times translates into letting the power switch device operate more in its active region, inevitably leading to increased losses. Consequently, in power electronics there are often a tradeoff between EMC performance and efficiency, while in digital signal systems the price to pay is instead mainly speed.

### 3.4.2 Spectrum Spreading

Just like rise and fall time control, spectrum spreading is a source suppression technique. A converter operated with a fixed frequency and duty cycle will exhibit a noise spectrum with several narrow harmonic peaks. By modulating the switching waveform in one way or another, the frequency content can be smoothed out or spread over a larger frequency range. Observe that the total energy of the signal can still be unaffected while at the same time the spectral peak content is decreased. For further reading, [21] can be a good start.

### 3.4.3 Snubbing

As mentioned in section 3.3 the ideal voltage square wave at the switch node in a buck converter will exhibit more or less ringing in any practical converter. The purpose of a snubber network is to limit these switching transients. There exist a variety of snubber circuits, where different combinations of the circuit elements, \( R, L, C \& D \) are employed. Snubber circuits with a resistive element will dissipate power, while snubbers without any resistors can be non-dissipative.

A snubber implementation quite often seen in a buck converter is a series combination of \( R \) and \( C \) placed parallel across the catch diode, or freewheeling
diode, which are both common names for the diode $D$ in Figure 1. Physically this snubber should be placed as close to the active switching device as possible rather than to the diode [22]. Such a snubber will however compromise the overall efficiency of the converter, and hence the design and use of a snubber should be carefully considered [23]. Except lower EMI, another benefit of less over and undershoot is that the voltage stress on the active switching device is reduced.

### 3.4.4 Filtering

The switching action of a buck converter makes it inherently noisy, and filtering of both input and output are therefore a necessity to keep noise from spreading throughout the system. Besides; the output should ideally be a pure DC voltage, and so any residue from the switching is highly unwanted. Since the input current of a buck converter is discontinuous, input filtering becomes even more important. A detailed analysis of input filtering can be found in [3]. A fact that can be worth to remember when it comes to filtering; is that common mode currents and differential mode currents generally require different filter strategies.

### 3.4.5 Shielding

Shielding is maybe one of the most commonly associated countermeasures when it comes to EMI problems. A shield is perhaps most often thought of as something that blocks electromagnetic fields, and while this is indeed often true, shielding can sometimes rely on quite different properties.

Shielding against magnetic fields can be such an example; a signal can be protected against magnetic interferences by the use of a shielded cable, provided it is grounded in both ends. Or components can be shielded from magnetic fields by high permeability materials, which divert magnetic fields rather than block them.

A shielded enclosure is often expensive and sometimes even unpractical, and therefore not the first choice when solving an EMI problem. The switch node in a buck converter is a source for capacitive coupling, to minimize this coupling a so called Faraday screen can be used, this is especially applicable if the switch node is made large, for example by the use of a heat sink. Extensive shielding theory can be found in [24].
4 Computer Simulations

In an attempt to characterize the CISPR 25 measurement setup seen in Figure 2, a number of 3D electromagnetic simulations were performed. CST MWS\textsuperscript{6} was the software used and a time domain based solver was employed.

4.1 Modeling of the Measurement Setup

To simulate the real measurement situation an aluminum plane with dimension according to Figure 17 was modeled. This plane will be referred to as ground plane throughout this chapter. The thickness of the ground plane was set to 10 mm. The monopole measurement antenna was modeled as a solid aluminum cylinder with radius 1 mm placed 5 mm above the ground plane.

Since the real measurement antenna is an active monopole, the bottom of the aluminum rod was loaded with a lumped element of 3 pF of capacitance in parallel with 1 M\Omega of resistance; to account for the input impedance of antenna amplifier.

\textsuperscript{6} CST STUDIO SUITE\textsuperscript{®} 2012, CST MICROWAVE STUDIO\textsuperscript{®}, Release Version 2012.05
The voltage over this lumped element was observed and if nothing else is stated this is the voltage referred to by the term *antenna voltage*. The origin marked (0, 0, 0) in Figure 17 will be used as a space reference in this chapter and corresponds to the coordinates \((x, y, z)\) expressed in meters.

### 4.2 Electric versus Magnetic Source

To investigate if the measured antenna voltage had any dependence on the source characteristic, simulations with an electric and magnetic source respectively were carried out. The electric source was represented by a short dipole—3 mm arms with a 2 mm gap—fed by a voltage source. The magnetic source was represented by a small—3 × 8 mm rectangle—loop fed by a current source. Both sources were modeled with perfect electric conducting wires.

#### 4.2.1 Electric Source in Free Space

The electric source (voltage source center) was placed at \((0.25, 0.75, 0.05)\), aligned with the Z-axis and excited with a voltage of 13 V. This simulation included only the electric source in free space, and the Z-directed electric far field was probed in \((0.25, 10, 0.05)\), which is plotted in Figure 18.

![Electric far field from electric source, Z-component](image)

As we can see, the electric field strength as a function of frequency is almost a straight line with a slope of 40 dB/decade. The slight deviation in the lower frequency region is most probably due to the fact that the bounding volume, in
which the field distribution was calculated was limited to save computation time; remember that the free space wavelength for a 150 kHz electromagnetic wave is 2 km.

Equation (2) for the far field of a Hertzian dipole predicts a field strength variation of only 20 dB/decade. A closer look at the simulation data reveals that the current from the voltage source grows with 20 dB per decade of increased frequency. This suggests that the small dipole antenna represents a mainly capacitive load. Knowing that the dipole current varies with 20 dB/decade, the field strength variation of 40 dB/decade in Figure 18 makes perfect sense, and we conclude that the simulation makes a decent representation of a small electric dipole.

### 4.2.2 Magnetic Source in Free Space

Here the magnetic source (loop center) was placed at (0.25, 0.7485, 0.05). The loop was oriented to lie in the Y–Z-plane and excited with a current of 10 A. As for the electric source simulation in section 4.2.1 this simulation also included only the magnetic source in free space, and the Z-directed electric far field was again probed in (0.25, 10, 0.05). Figure 19 shows a plot of that probed field.

![Figure 19. Electric far field from magnetic source, Z-component](image)

Again we see that the plotted field strength in a large frequency range resembles a straight line with a slope of 40 dB/decade, which is consistent with equation (9) which describes the maximum far field strength from an electrically small loop. The deviation from the straight line in the lowest part of the frequency range may
be explained by the same reasons stated in section 4.2.1 for the electric source. Note that the current and voltage were deliberately chosen so as to produce nearly the same far field strength by both the electric and magnetic source.

4.2.3 Electric & Magnetic Source with Measurement Setup

Now the structure depicted in Figure 17 was added to the model and with everything else unchanged the simulations described in section 4.2.1 and 4.2.2 were repeated. The distance from the source to the center of the monopole is now roughly 1.5 m. So for a strictly 1/r dependence and no field influence from the measurement arrangement, and without account to any antenna factor, we would expect the antenna voltage to simply reflect the plots of Figure 18 and Figure 19 but with an offset in the range of 16.5 dB (20log(10/1.5)). Looking at Figure 20 we see a complete different set of curves.

![Antenna Voltage vs. Frequency](image)

**Figure 20.** Antenna voltage for electric and magnetic source respectively

First of all, none of the curves has a slope near 40 dB/decade. Trace \( T\# [1] \) corresponds to the electric source and \( T\# [2] \) to the magnetic source. Furthermore we note that at 30 MHz the two curves approaches each other and we are not too far away from the above far field reasoning that would give us an antenna voltage magnitude of -53.5 dB (-70+16.5).

Recall that a frequency of 30 MHz corresponds to a free space wavelength of 10 m, indicating that a distance of 1.5 m is indeed approaching the region where we can start to think about using far field approximations.
But for the majority part of the frequency range; 150 kHz – 30MHz the two traces not only deviate significantly from the far field expressions, but also do they greatly differ between themselves. At the lower end of the frequency span the difference is more than 50 dB, and while the electric source produces a nearly constant antenna voltage with respect to frequency, the antenna voltage emanating from the magnetic source varies as 20 dB/decade.

In order to try to get some understanding of these differences, we turn our attention to the full set of equations describing the complete electric, and magnetic field distribution around a Hertzian dipole and an electrically small loop. These equations are derived from the fundamental relationships described by the set of equations known as Maxwell’s equations.

The derivation of the former mentioned equations can be found in many decent textbooks on antennas or electromagnetic field theory, such as [8] or [25]. According to [25], the vector components of the magnetic and electric field around an infinitesimal dipole carrying a uniform time harmonic current, can be expressed as:

\[
H_r = 0
\]  
\[
H_\theta = 0
\]  
\[
H_\phi = j \frac{\beta_0 \beta_0}{2\pi r} \sin \theta \left[ 1 + \frac{1}{j \beta_0 r} \right] e^{-j\beta_0 r}
\]
\[
E_r = j \frac{\eta_0 \beta_0}{2\pi r} \cos \theta \left[ 1 + \frac{1}{j \beta_0 r} + \frac{1}{(j \beta_0 r)^2} \right] e^{-j\beta_0 r}
\]
\[
E_\theta = j \frac{\eta_0 \beta_0}{2\pi r} \sin \theta \left[ 1 + \frac{1}{j \beta_0 r} + \frac{1}{(j \beta_0 r)^2} \right] e^{-j\beta_0 r}
\]
\[
E_\phi = 0
\]

where

\[
\eta_0 = \text{Intrinsic impedance of free space} \approx 120\pi \ \Omega
\]
\[
\beta_0 = \text{Free space wavenumber} = \frac{2\pi}{\lambda_0} \ \text{[m}^{-1}\text{]}
\]
\[
\lambda_0 = \text{Free space wavelength} \ [\text{m}] 
\]
\[ \dot{I} = \text{Current amplitude} \ [\text{\AA}] \]

\[ dL = \text{Length of dipole} \ [\text{m}] \]

\[ r = \text{Distance} \ [\text{m}] \]

\[ \theta = \text{Polar angle} \]

and the dipole is positioned at the origin and along the radial direction when the polar angle \( \theta \) equals zero, in a spherical coordinate system described by the orthogonal unit vector triple; \( \hat{a}_r, \hat{a}_\theta, \hat{a}_\phi \), which point in the radial direction, polar angular direction and azimuth angular direction respectively. Employing the same coordinate system, the vector components describing the field surrounding an electrically small circular loop carrying a uniform time harmonic current, placed at the origin and in the \( r-\Phi \)-plane when the polar angle equals \( \pi/2 \) radians, can be expressed as [25]:

\[ E_r = 0 \quad \text{(16)} \]

\[ E_\theta = 0 \quad \text{(17)} \]

\[ E_\phi = \frac{\beta_0 \omega \mu_0 \dot{I}}{4 \pi r} \sin \theta \left[ \frac{1}{j \beta_0 r} + \frac{1}{(j \beta_0 r)^2} \right] e^{-j \beta_0 r} \quad \text{(18)} \]

\[ H_r = -\frac{\beta_0 \omega \mu_0 \dot{I}}{2 \pi \eta_0 r} \cos \theta \left[ \frac{1}{j \beta_0 r} + \frac{1}{(j \beta_0 r)^2} \right] e^{-j \beta_0 r} \quad \text{(19)} \]

\[ H_\theta = -\frac{\beta_0 \omega \mu_0 \dot{I}}{4 \pi \eta_0 r} \sin \theta \left[ 1 + \frac{1}{j \beta_0 r} + \frac{1}{(j \beta_0 r)^2} \right] e^{-j \beta_0 r} \quad \text{(20)} \]

\[ H_\phi = 0 \quad \text{(21)} \]

where

\[ \eta_0 = \text{Intrinsic impedance of free space} \approx 120\pi \ [\Omega] \]

\[ \beta_0 = \text{Free space wavenumber} = \frac{2\pi}{\lambda_0} \ [\text{m}^{-1}] \]

\[ \lambda_0 = \text{Free space wavelength} \ [\text{m}] \]

\[ \dot{I} = \text{Current amplitude} \ [\text{\AA}] \]

\[ A = \text{Area of loop} \ [\text{m}^2] \]
\[ \omega = \text{Angular frequency [s}^{-1}\text{]} \]

\[ \mu_0 = \text{Permeability of free space [H/m]} \]

\[ r = \text{Distance [m]} \]

\[ \theta = \text{Polar angle} \]

The wave impedance is defined as the ratio between the electric and magnetic fields. Using equations (10)–(15) and (16)–(21) we can calculate the wave impedance for an electrical source and magnetic source respectively. The magnitude of the wave impedances when the polar angle \( \theta \) equals \( \pi/2 \) radians were calculated for distances up to \( 0.5 \lambda_0 \), and a plot of these are shown in Figure 21. The dashed line indicates the value of the intrinsic impedance of free space: approximately \( 377 \ \Omega \).

![Figure 21. Wave impedance magnitude, electric and magnetic source](image)

Looking at Figure 21 we see that compared to the intrinsic impedance of free space, the electromagnetic field close an electric source is totally dominated by the electric field, whereas close to a magnetic source, the opposite is true. At larger distances both traces seem to converge toward \( 377 \ \Omega \), which is precisely what we expect.
The fact that the wave impedances in the near field are completely different, although the far field wave impedances are equal explains the discrepancy we see in Figure 20. The simulation data for the current loop shows that the voltage at the current source feeding the loop increases virtually as 20 dB/decade with frequency\(^7\), while for the Hertzian dipole the voltage at the source is basically constant. The monopole antenna is in essence an electric field sensor and from Figure 20 we thus conclude that in the near field region, it indeed seems to be the source voltage that is the determining factor for the antenna voltage. Again remember that the two sources produce virtually the same electric field strength in the far field region.

4.3 Emissions from Cables

Looking back at the test setup in Figure 2 we see that both the power supply and possible load are connected to the equipment under test through cables; these cables are specified to be roughly 1.5 meters long. To investigate how any voltage or current on these cables may contribute to the radiated emissions, a few simulations were run. Again the basic structure in Figure 17 was used. This time a pair of 1.5 m long wires were added, positioned according to Figure 2. The distance between the two wires was set to 5 mm.

4.3.1 Voltage Source

In this simulation a voltage source, \(V_S\) was connected to one end of the wires and the other end was loaded by a resistive load, \(R_L\). Different source voltage and load resistance combinations were simulated according to Table 3.

<table>
<thead>
<tr>
<th>#</th>
<th>(V_S) [mV]</th>
<th>(R_L) [Ω]</th>
<th>(I_{DC} = V_S/R_L) [µA]</th>
<th>(V_{S}/V_{S#1}) [dB]</th>
<th>(I_{DC}/I_{DC#1}) [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>10</td>
<td>100</td>
<td>100</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>2</td>
<td>20</td>
<td>200</td>
<td>100</td>
<td>6</td>
<td>0</td>
</tr>
<tr>
<td>3</td>
<td>30</td>
<td>300</td>
<td>100</td>
<td>9.5</td>
<td>0</td>
</tr>
<tr>
<td>4</td>
<td>40</td>
<td>400</td>
<td>100</td>
<td>12</td>
<td>0</td>
</tr>
<tr>
<td>5</td>
<td>100</td>
<td>1000</td>
<td>10</td>
<td>60</td>
<td>40</td>
</tr>
</tbody>
</table>

Table 3. Simulation data for Figure 22, cable fed by voltage source

---

\(^7\) For a constant current, a 20 dB/decade increase in voltage with frequency is consistent with a pure inductive load.
The last two columns in this table indicate the ratio (in dB) of voltage and current relative the values in the first row respectively. A plot of the simulated antenna voltages can be seen in Figure 22.

![Antenna Voltage vs Frequency](image)

**Figure 22. Emissions from cables fed by a voltage source**

In this plot we see a set of traces that appears to be vertically translated. A closer look reveals that the offset from T\# [1], very closely matches the source voltage ratio tabulated in Table 3. The frequency response is basically flat up to approximately 10 MHz, where the antenna voltage begins to increase with frequency. At higher frequencies the voltage amplitude between the wires can no longer be considered constant along the wires.

A look at the simulated voltage at the load end reveals that for all the loads in Table 3 the voltage at this end tends to increase with frequency while the source voltage is constant over the entire range. This fact partly explains the frequency dependency we see at higher frequencies. Adding 120 dB to the levels in Figure 22 gives the levels in dB(µV), and we see that a few tens of mV between the wires are enough to potentially cause EMC compliance trouble.

From transmission line theory we know that the standing wave ration, SWR is a function of the characteristic impedance of the line and the load impedance. The locations of the maxima and minima of a voltage standing wave are also affected by the ratio of characteristic and load impedances. For a generator matched and lossless transmission line with characteristic impedance (resistance), $Z_0$ and a
purely resistive load, $R_L$, the voltage standing wave pattern will exhibit either a maximum or minimum at the load end; $R_L > Z_0$ yields a maximum while $R_L < Z_0$ gives a minimum. For a matched case where $R_L = Z_0$ the voltage standing wave ratio (VSWR) equals unity and there are neither maxima nor minima. To explore this theory another simulation with two different load resistances was carried out. Figure 23 shows a plot of the results. The source voltages in both cases were set to 10 mV. Table 4 presents a mapping for the different traces.

<table>
<thead>
<tr>
<th>$R_L$ [Ω]</th>
<th>Source voltage</th>
<th>Load voltage</th>
<th>Antenna voltage</th>
</tr>
</thead>
</table>

Table 4. Trace map for Figure 23

As we see in Figure 23 the source voltage in both cases are constant. For a load resistance of 100 Ω the voltage at the load end increases with frequency whereas when the load resistance is 15 Ω the load voltage decreases with frequency.

This behavior is consistent with the transmission line theory provided that the characteristic impedance of two wire cable is somewhere between 15 Ω and 100 Ω. If we look at the antenna voltages we see that T# [5] increases more with frequency than the corresponding load voltage. T# [6] is almost flat with a slight tendency to increase at the end of the frequency range while the corresponding load voltage is actually decreasing with frequency.
This suggests the existence of another mechanism that affects the antenna voltage. This could be characterized as a transfer function of the measurement setup itself, which would be the transfer function from cable voltage to antenna voltage, which in this specific case seem to increase slightly with frequency. The antenna voltage is then a combination of the cable voltage and this transfer function.

4.3.2 Current Source

Now, the voltage source was replaced by a current source. The different source current and load resistance combinations that were simulated are as follows:

<table>
<thead>
<tr>
<th>#</th>
<th>$R_L$ [$\Omega$]</th>
<th>$I_S$</th>
<th>$V_{DC} = R_L \times I_S$</th>
<th>$I_S/I_{S#1}$ [dB]</th>
<th>$V_{DC}/V_{DC#1}$ [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>10</td>
<td>1 A</td>
<td>10 V</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>2</td>
<td>5</td>
<td>2 A</td>
<td>10 V</td>
<td>6</td>
<td>0</td>
</tr>
<tr>
<td>3</td>
<td>3.33</td>
<td>3 A</td>
<td>10 V</td>
<td>9.5</td>
<td>0</td>
</tr>
<tr>
<td>4</td>
<td>2.5</td>
<td>4 A</td>
<td>10 V</td>
<td>12</td>
<td>0</td>
</tr>
<tr>
<td>5</td>
<td>1</td>
<td>10 mA</td>
<td>10 mV</td>
<td>-40</td>
<td>-60</td>
</tr>
</tbody>
</table>

Table 5. Simulation data for Figure 24, cable fed by current source

The simulated antenna voltages are plotted in Figure 24. This time we see a different behavior within the set of traces compared to the previous case.

![Figure 24. Emissions from cables fed by a current source](image)

At frequencies up to a few MHz, the antenna voltage is almost constant and as can be seen from $T# [1]-[4]$, the antenna voltage level is virtually independent of the
current source magnitude. Furthermore, in the low frequency region, $T^# [5]$ is offset from $T^# [1-4]$ by the DC-voltage ratio rather than the current ratio, see Table 5.

Since the wires were fed by a current source the inductance of the circuit will cause the voltage at the source end to increase with frequency. This fact will make the antenna voltage to increase with frequency as well. As we can see in Figure 24 this effect is more pronounced for lower load resistances, this is because the impedance contribution from the circuit inductance, makes up for more and more of the total circuit impedance for lower load resistances. The same explanation about a varying voltage along the cable in the voltage source fed case applies here as well. Keep in mind that for a typical buck converter switching frequency of a few hundred kHz, the fundamental plus the ten or so first harmonics all lies in the frequency range where the traces in Figure 24 are fairly flat.

### 4.4 Evaluation of a Simple Lumped Element Model

All simulations this far indicates that it is voltage and not current that is the decisive quantity for the antenna voltage in a CISPR 25 measurement setup. For low frequencies where the structure is electrically small, electronic circuits are successfully modeled by lumped circuit elements. The simulations described in this section are intended to evaluate a simple lumped element model of the CISPR 25 setup and a voltage fed structure. A capacitive coupling model according to Figure 25 is proposed.

![Capacitive coupling model](image)

**Figure 25. Capacitive coupling model**
In this figure, 1 and 2 represent two galvanically isolated metal structures and 3 represents the measurement antenna. If we connect a voltage source between 1 and 2, the model can be represented by the schematic in Figure 26.

![Figure 26. Lumped model schematic](image)

In this schematic the mutual capacitance between 1 and 2 has been omitted and the voltage over C3 would be the antenna voltage. Furthermore we see that if the ratio between C13 and C23 equals the ratio between C1 and C2 the antenna voltage becomes zero. Again, for simulation, the structure in Figure 17 was used and a short dipole identical to the one described in section 4.2 was placed in different locations.

This time the dipole was aligned with the X-axis; observe that this orientation makes the short dipole and the measurement antenna perpendicular to each other, and so are their far field E-planes. The dipole was fed by a voltage source of 1 V and first placed with its center in (1, 0.9, 0.05). With this placement the two halves of the dipole are positioned completely symmetric with respect to the measurement setup, and the closest distance to the measurement antenna is 1 m. The simulated antenna voltage for this setup can be seen as $T\# [1]$ in Figure 27.
Next the dipole was moved to (0.7, 0.9, 0.05), this moves the dipole 30 cm off the symmetry line and the distance to the measurement antenna becomes slightly longer rather than shorter. $T^\#[2]$ in Figure 27 corresponds to the simulated antenna voltage for this placement. From this figure we see that the movement of the dipole by 30 cm off the symmetry line, resulted in an increase of the antenna voltage in the low frequency region by approximately 75 dB or more than 5600 times. This suggests that the model in Figure 26 may be a good representation of the situation at lower frequencies. This result also demonstrates the complexity of the near field and why distance dependent approximations are not feasible.

In Figure 27 we see that $T^\#[2]$ is roughly frequency independent up to approximately 10 MHz, where the wavelength becomes short enough compared to the distances in the measurement setup, to invalidate the lumped model. Here $T^\#[1]$ does not exhibit the same frequency dependency, which could be because of the fact that the simulation has simply “bottomed out” at -200 dB.

### 4.5 Models of Parts of a Buck Converter

In this section a few different geometries which could be representative for different parts of a buck converter, will be simulated. As for all the simulations in this chapter; even though we get a result in terms of an antenna voltage, we do not consider the absolute value of this voltage as the primary result from the simulations, but rather how different parameters affects this antenna voltage.
Since there are many parameters that affect the antenna voltage and small differences in placement can have a huge impact, the actual antenna voltage in a real measurement may differ a lot from the simulated value while relative trends are still hopefully coherent.

### 4.5.1 A Voltage Switch Node

We begin with a square patch of copper placed over a larger square of copper sheet. This would be a good representation of a switch node over a reference plane in a buck converter. The thickness of the square copper patches was set to 35 µm and the distance between them was set to 1.6 mm, these values are typical for a PCB. A voltage source was then placed in the center between the two patches.

Figure 28 shows the dimensions of the structures used for these simulations. These structures were then integrated with the model of Figure 17. The center point of the underside of the larger patch was located at (0.25, 0.75, 0.05). First structure a in Figure 28 without the screen was simulated with different sizes of the switch node according to Table 6.

![Figure 28. Geometry of the simulated switch node](image)

<table>
<thead>
<tr>
<th>#</th>
<th>Source voltage [V]</th>
<th>Patch side, a [mm]</th>
<th>Patch area, A [mm²]</th>
<th>20log(A#/A#-1)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>24</td>
<td>5</td>
<td>25</td>
<td>N/A</td>
</tr>
<tr>
<td>2</td>
<td>24</td>
<td>10</td>
<td>100</td>
<td>12</td>
</tr>
<tr>
<td>3</td>
<td>24</td>
<td>15</td>
<td>225</td>
<td>7</td>
</tr>
<tr>
<td>4</td>
<td>24</td>
<td>20</td>
<td>400</td>
<td>5</td>
</tr>
<tr>
<td>5</td>
<td>24</td>
<td>25</td>
<td>625</td>
<td>4</td>
</tr>
</tbody>
</table>

Table 6. Simulation parameters for T# [1] – [5] in Figure 29
Plotted antenna voltages for these simulations can be seen in Figure 29. Here we see that a larger switch node gives a higher antenna voltage.

![Figure 29. Antenna voltage for the structure in Figure 28a](image)

With aid of the last column in Table 6 we can see that in this case the antenna voltage seems to be roughly proportional to the switch node area. But, by this observation not said that this is a general behavior. Again we see an almost flat frequency response in the lower to mid frequency range. $T\# [6]$ is the result of a simulation with the same parameters as for $T\# [1]$, except that a dielectric slab with thickness 1.6 mm and the same size as the larger square was inserted between the copper squares. The dielectric was modeled as FR4 (loss free) with dielectric constant, $\varepsilon_r=4.3$. From Figure 29 we see that the addition of a dielectric lowered the antenna voltage. The reason for this decrease is that a higher permittivity concentrates the electric field to in between the copper patches and so we have a less fringing field that can couple to the antenna.

After this simulation the dielectric was removed and the *screen* in structure a in Figure 28 was added. The antenna voltage from this simulation is presented as $T\# [7]$ in Figure 29. Compared to $T\# [1]$ the addition of the *screen* more than halved the antenna voltage.
Figure 30. Absolute value of E-field around switch node

Figure 30 is a graphic representation of the absolute value of the electric field around the switch node in the (X, 0.75, Z)-plane, a and b shows the field pattern without and with the screen respectively. From these pictures we can see the electrical shielding effect of the screen.

One way to think of the shielding effect is that with the screen in place, “fewer electrical field lines” are allowed to terminate at the antenna. Observe that the screen does not have to be directly in between the patch and antenna to “prevent some field lines” from reaching the antenna, but can as well be placed “behind” the patch, and still “steal some field lines”. A screen of the type in structure a in Figure 28 may not be very practical so a copper area according to structure b in Figure 28 was added and simulated as well. This copper area was either connected to the lower patch by two wires at the points marked “Connection points” in Figure 28 or left unconnected, in Table 7 this corresponds to “Screen Floating”.

Up until now, no connection between the simulated structure and the ground plane of the measurement setup has been present. To investigate how the antenna voltage may be affected by any grounding, the lower copper patch was grounded through an impedance in some of the simulations this time. The grounding was made by a lumped element from the center of the lower copper square to the ground plane beneath. This impedance was configured as a series combination with 50 Ω resistance and 100 nF capacitance. This combination was chosen based on a typical LISN used in this type of measurements. This grounding corresponds to “Grounded” in Table 7 which shows the varying simulation parameters.
Table 7. Simulation parameters for the traces in Figure 31

<table>
<thead>
<tr>
<th>#</th>
<th>Screen</th>
<th>Screen floating</th>
<th>Grounded</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>No</td>
<td>N/A</td>
<td>No</td>
</tr>
<tr>
<td>2</td>
<td>Yes</td>
<td>Yes</td>
<td>No</td>
</tr>
<tr>
<td>3</td>
<td>Yes</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td>4</td>
<td>No</td>
<td>N/A</td>
<td>Yes</td>
</tr>
<tr>
<td>5</td>
<td>Yes</td>
<td>Yes</td>
<td>Yes</td>
</tr>
<tr>
<td>6</td>
<td>Yes</td>
<td>No</td>
<td>Yes</td>
</tr>
</tbody>
</table>

The switch node was excited by a voltage source with a voltage level of 24 V. The dimensions of the structure can be seen in Figure 28. The simulated antenna voltages are shown in Figure 31.

![Figure 31. Antenna voltage for the structure in Figure 28b](image)

*T# [1]* in this figure is merely the result of repeated simulation of the same structure corresponding to *T# [1]* in Figure 29. From *T# [2]* we see that the addition of the copper area around the switch node increases the emissions when it is left floating. If this copper area is electrically connected to the larger copper square, the antenna voltage instead decreases as we see from *T# [3]*. *T# [4]–[6]* is basically *T# [1]–[3]* shifted approximately 4 dB. This means that the grounding of the structure increases the antenna voltage with equal amount regardless of screen configuration. From the simulations in this section we can draw at least one important conclusion; the radiated emissions from a switch node can be lowered by adding a conductive screen in the proximity of the switch node.
It is equally important to remember that this screen should be connected to a fixed potential, because as we have seen, adding a screen and leave it floating can actually worsen the emissions.

4.5.2 A Vertical Switch Current Loop

As described previous in this text, the input current to a buck converter is inherently discontinuous. This makes pulsating current loops unavoidable in buck converter designs. To investigate the properties of radiated emissions from such current loops, a few simulations were performed. The models used in these simulations were built on the same structure and placement as described in section 4.5.1 and depicted in Figure 28. First the switch node was removed and replaced by a wire loop and current source according to Figure 32.

![Figure 32. Model for vertical loop simulations](image)

A constant current was used and the loop dimensions were altered according to Table 8.

<table>
<thead>
<tr>
<th>#</th>
<th>Source current [A]</th>
<th>Loop length, a [mm]</th>
<th>Loop height, h [mm]</th>
<th>Loop area, a × h [mm²]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>20</td>
<td>1.6</td>
<td>32</td>
</tr>
<tr>
<td>2</td>
<td>1</td>
<td>20</td>
<td>3.2</td>
<td>64</td>
</tr>
<tr>
<td>3</td>
<td>1</td>
<td>40</td>
<td>1.6</td>
<td>64</td>
</tr>
<tr>
<td>4</td>
<td>1</td>
<td>40</td>
<td>3.2</td>
<td>128</td>
</tr>
</tbody>
</table>

Table 8. Simulation parameters for the traces in Figure 33 and Figure 34
As can be seen in Figure 33 the antenna voltages are basically straight lines with slopes of 20 dB/decade.

![Antenna Voltage Graph](image1)

**Figure 33. Antenna voltage from vertical loop simulations**

We also note that $T\# [3]$ and $T\# [4]$ are almost identical; these two traces correspond to the same loop area but with different proportions. From Figure 33 and Table 8 we also see that the offset between the traces pretty much relates to the loop area ratio. The source voltages from the simulations are presented in Figure 34.

![Source Voltage Graph](image2)

**Figure 34. Source voltage from vertical loop simulations**
This figure will help us understand the frequency dependency of the antenna voltage. Even though the current through the loop is constant over the whole frequency range, Figure 34 reveals that the voltage at the current source increases with 20 dB/decade and is also approximately proportional to the loop area. This behavior of the source voltage is consistent with a purely inductive load. Yet again we conclude that it seems to be voltage rather than current which is the determining factor for the antenna voltage. Even if it is voltage that determines the radiated emissions, current and current loops cannot be neglected because of the simple fact that the voltage across a loop inductance, increases proportionally with both current and inductance as well as with frequency, where inductance increases with loop area.

In Figure 34 we also see that the traces are straight lines over the whole frequency range whereas in Figure 33 the traces tend to bend up slightly for higher frequencies. The interpretation of this is that at 30 MHz the loop dimensions are still electrically sufficiently small while the dimensions of the measurement setup are large enough to affect the result for the antenna voltage.

To see the effect of different loop currents, the loop with dimensions according to the first row in Table 8 was reused and simulated with different current levels. Four different current magnitudes were simulated: 1, 2, 3 and 4 amperes, these current levels correspond to $T^# [1], [2], [3]$ and $[4]$ respectively in both Figure 35 and Figure 36. In these figures we see a similar set of traces as for the different loop areas previously simulated. But in this case it is the different source current levels that cause the offsets rather than different inductances of the loops. In both Figure 35 and Figure 36 we see that the offsets correspond very well to the different current levels.

---

8 Remember that the inductance of a loop is not necessarily proportional to the loop area.
4.5.3 A Lateral Switch Current Loop

Another possible geometry of a switch loop is a loop parallel to a reference plane. Such a conducting plane is sometimes called an image plane [26] [27] [28]. Figure 37 shows the model of the loop used in these simulations, one corner of the loop was electrically connected to the reference plane, this because that was thought to be the most realistic situation.
First, both loop dimensions and distance from the reference plane were altered according to Table 9, while the current through the loop was held constant.

<table>
<thead>
<tr>
<th>#</th>
<th>Source current [A]</th>
<th>Loop length, a [mm]</th>
<th>Loop width, w [mm]</th>
<th>Loop height, h [mm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>20</td>
<td>1.6</td>
<td>1.6</td>
</tr>
<tr>
<td>2</td>
<td>1</td>
<td>20</td>
<td>1.6</td>
<td>3.2</td>
</tr>
<tr>
<td>3</td>
<td>1</td>
<td>40</td>
<td>3.2</td>
<td>1.6</td>
</tr>
<tr>
<td>4</td>
<td>1</td>
<td>40</td>
<td>3.2</td>
<td>3.2</td>
</tr>
</tbody>
</table>

Table 9. Simulation parameters for the traces in Figure 38 and Figure 39

A plot of the simulated antenna voltages can be found in Figure 38.
In this figure we see that the antenna voltage displays the same behavior as for the vertical loop. The levels increase with loop area and also with the height of the loop above the reference plane. Figure 39 shows a plot of the simulated voltage at the current source feeding the loop.

![Figure 39: Source voltage from lateral loop simulations](image)

Here we see a small increase in voltage when the loop is placed higher above the reference plane, indicating that a conducting plane close to a loop lower its effective inductance. This time though the relative differences between the traces in Figure 38 and the traces in Figure 39 are not similar, so we conclude that the geometry of the different loops also significantly affected the antenna voltage. However, the behavior of the antenna voltage reflects the behavior of the corresponding source voltage and not source current, which is constant over the whole frequency range.
As for the vertical loop, the loop with the dimensions according to the first row of Table 9 was now simulated with different current magnitudes. The simulated antenna voltage and source voltage can be seen in Figure 40 and Figure 41 respectively.

![Antenna Voltage](image1)

**Figure 40. Antenna voltage for different lateral loop current magnitudes**

![Source Voltage](image2)

**Figure 41. Source voltage for different lateral loop current magnitudes**

\(T\# [1], [2], [3] \text{ and } [4]\) in both of these figures correspond to a source current of 1, 2, 3, and 4 amperes respectively.
This time we see a good correspondence in relative differences for the antenna voltage and source voltage plots. The offsets also follow the current ratio very well. By now, this is precisely what we would expect, knowing that the geometry is the same for all traces.

4.5.4 A Switch Current Loop with Attached Cables

Since radiation from cables attached to printed circuit boards are a known issue, see for example [16], the models used in the simulations described in section 4.5.2 and 4.5.3 were rebuilt with an added wire and simulated again. A 1.5 m long wire was attached to the reference plane and extended in the X-direction, which is consistent with the CISPR 25 measurement setup. The results from these simulations all indicated the same thing: the added wire had very little influence on the antenna voltage. This indicates that there are other mechanisms dominating the antenna voltage so that any current on the added wire did not affect the overall result.
5 Radiated Emission Measurements

To practically investigate how different parameters in a buck converter design impact on radiated emission measurements, a number of test boards were developed. The radiated emissions from these test boards were then measured using a setup similar to the CISPR 25 measurement setup. It should be noted here that the test setup is in no way accredited or calibrated, thus all measurement results presented here shall only be used for relative comparisons, which for our purpose is fully satisfactory.

5.1 Test Setup

The radiated emission measurements were performed using the measurement setup described in Appendix A. The physical arrangement of the test setup is depicted in Figure 42.

![Figure 42. Measurement setup](image)

The battery was placed on the floor in the shielded enclosure and the rest of the equipment in Figure 42 was located on the metal table. The device under test, DUT, resistive load and cables on the table were placed on a 50 mm thick foam
sheet. As will be seen later some of the measurements were performed with a shielded load cable and or power supply cable. In these cases the respective cable was placed in a metal tube which was then connected to and laid directly onto the metal table. Both the load and supply cables were made up by twisted AWG #20 wires. AWG #20 corresponds to a wire cross section area of approximately 0.5 mm$^2$.

### 5.2 The Test Boards

As test circuit a buck converter built around the TPS54260 [29] was employed. The TPS54260 is a 60 V, 2.5 A, step down regulator IC from Texas Instruments. The switching frequency of the buck converter was set to slightly above 300 kHz, and the output voltage was set to 3.3 V. All the test boards used for the measurements described in this chapter were built from the same electrical schematic. The complete schematic can be found in Appendix B. The test boards instead differ through varying board layouts, including utilization of the PCB stack-up. A total of five different layouts were designed, some of which allowed for further variant configuration. A component placement drawing that is valid for all five of the test boards is to be seen in Appendix D. The input filter components; C1–C3, C11 & L1 were not mounted, L1 being replaced by a solid wire jumper. All the test boards were manufactured with the same PCB stack-up, which is described in Figure 43.

![Figure 43. Test PCB build-up](image)

Notice that the Cu-foil thickness on the outer layers is given as base copper thickness$^9$. We will now take a closer look at each of the individual test board and associated measurements.

---

$^9$ Typically additional 25-30 µm of copper will be added through plating.
5.2.1 Test Board 1

A picture of an unpopulated PCB for test board 1, TB1 is presented in Figure 44. The copper areas marked 1, 2 & 3 in Figure 44 are connected to $U1:10$ in the schematic and provide a way to alter the area of the switch node. Information and images from the production files of all four copper layers are to be found in Appendix E.

![Figure 44. Test board 1](image)

A buck converter of the type described in chapter 1 has as mentioned two\textsuperscript{10} basic modes of operation. To make the measured emission levels from different measurements to be relatively comparable to each other, we need the converter to operate in the same mode for all the measurements. The load resistance at the limit, $R_{lim}$ between continuous and discontinuous conduction mode can be expressed as [3]:

$$R_{lim} = 2f_sL \frac{V_{in}}{V_{in}-V_{out}} \quad (22)$$

Using a switching frequency, $f_s$ of 300 kHz, an inductor value, $L$ of 10 $\mu$H, an input voltage of 24 V and an output voltage of 3.3 V, equation (22) yields a resistance of

\textsuperscript{10} To further increase efficiency at light loads many buck controllers incorporates a third mode which typically implements some type of pulse skip or burst operation.
7 \Omega. Thus to ensure that the converter operates in continuous conduction mode the load resistance shall not exceed 7 \Omega. First a set of measurements according to Table 10 was made.

<table>
<thead>
<tr>
<th>#</th>
<th>Supply voltage [V]</th>
<th>Load resistance [Ω]</th>
<th>Shielded supply cable</th>
<th>Shielded load cable</th>
<th>Extra switch node area</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>23.5</td>
<td>2.35</td>
<td>No</td>
<td>No</td>
<td>1+2+3</td>
</tr>
<tr>
<td>2</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>1+2+3</td>
</tr>
<tr>
<td>3</td>
<td>23.5</td>
<td>2.35</td>
<td>Yes</td>
<td>Yes</td>
<td>1+2+3</td>
</tr>
</tbody>
</table>

Table 10. Measurement parameters for the traces in Figure 45

The results from those measurements are presented in Figure 45.

Figure 45. Radiated emissions measurements according to Table 10

In this figure we can see that there is no significant difference between the different traces. Because of this we conclude that neither the magnitude of the load current nor emissions from the cables are the dominant cause for the radiated emissions in this case.

Next is a set of measurements intended to investigate the significance of the switch node area. Looking at the copper layers in Appendix E we can see how the area of the switch node at the top layer can be altered by simply cutting away traces on the PCB. Whenever the switch node area was decreased in this way, the remaining isolated copper island was soldered to the surrounding ground plane.
The openings in the solder mask around the switch node seen in Figure 44 were added to facilitate this grounding. The traces in Figure 46 show the measured radiated emissions according to Table 11.

![Figure 46. Radiated emissions measurements according to Table 11](image)

<table>
<thead>
<tr>
<th>#</th>
<th>Supply voltage [V]</th>
<th>Load resistance [Ω]</th>
<th>Shielded supply cable</th>
<th>Shielded load cable</th>
<th>Extra switch node area</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>1+2+3</td>
</tr>
<tr>
<td>2</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>2+3</td>
</tr>
<tr>
<td>3</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>3</td>
</tr>
<tr>
<td>4</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>–</td>
</tr>
</tbody>
</table>

Table 11. Measurement parameters for the traces in Figure 46

This time we see a clear difference between the different traces. As the area of the switch node is decreased the measured levels of radiated emissions go down. Note that at the fundamental switching frequency and the first few harmonics the difference between \( T\# [1] \) and \( T\# [4] \) is more than 10 dB.

To see if the orientation of the buck inductor has any substantial impact on the radiated emissions, \( L2 \) was rotated 180°. All other parameters being the same as for \( T\# [4] \) in Table 11 no differences could be detected between the different inductor orientations. Finally the supply voltage to the buck converter was lowered to see how this might affect the radiated emission measurements.
But before we take a closer look at the results let us see how changing the supply voltage affects the operation of the buck converter. As is evident from equation (1) in chapter 1 the duty cycle of the step down converter in continuous conduction mode is dependent on the input voltage.

Since the spectral content of a rectangle pulse train is heavily dictated by the pulse width to period ratio we have to be a little cautious with our conclusions when we compare the radiated emissions for different input voltages. Figure 47 is a snapshot of an oscilloscope measurement of the wave form at the switch node.

Figure 47. Switch node waveform with 23.5 V supply

The supply voltage in this case is 23.5 V. As we see from the oscilloscope trace, the duty cycle is 16 % and the switching frequency is 313.5 kHz. When the input voltage is lowered to 12.7 V the wave form at the switch node changes according to Figure 48, note the different scaling of the voltage axis.
From this picture we see that the switching frequency is still 313.5 kHz but the duty cycle has now increased to 30.5 %. These two waveforms obviously have different frequency content, a fact we must pay attention to when we are comparing the radiated emissions for the different supply voltages. To help draw any sensible conclusions from the measurements, we again turn to Fourier analysis and equation (4) in chapter 2. Using this equation and taking the absolute values, Table 12 was computed.

| n  | A = |a₀| × 16%| B = |a₀| × 30.5%| C = B/A [dB] | C × 12.7/23.5 [dB] |
|----|-----|------|------|------|-------------|------------------|
| 1  | 0.307 | 0.521 | 4.600 | -0.745 |
| 2  | 0.269 | 0.299 | 0.940 | -4.405 |
| 3  | 0.212 | 0.056 | -11.555 | -16.900 |
| 4  | 0.144 | 0.101 | -3.043 | -8.388 |
| 5  | 0.075 | 0.127 | 4.589 | -0.756 |
| 6  | 0.013 | 0.054 | 12.174 | 6.828 |

Table 12. Fourier coefficients and comparison

The last column in this table shows the relation between the two supply voltages, in dB, adjusted with the difference in frequency content. With this knowledge in mind we now take a look at the traces in Figure 49, which are measurements with parameters according to Table 13.
At first glance there appears to be no obvious correlation between the two traces. But if we look at the numbers in the last column of Table 12 and compare these with the differences between the traces in Figure 49, we see that there is actually a very good agreement between the differences in measured radiated emission and the calculations in Table 12.

In my opinion, this agreement is so good that it leaves little to no doubt that in this case the measured radiated emissions are completely dominated by the voltage level at the switch node.
5.2.2 Test Board 2

As can be seen in Figure 50, test board 2, TB2 appears to be identical to TB1. But looking in Appendix F we see that the ground plane is now placed on layer 2 instead of layer 4 as was the case for TB1.

![Figure 50. Test board 2](image)

Looking at the build-up in Figure 43 we see that this means that the ground plane in TB2 is more than 1 mm closer to the top layer with the switch node compared to TB1. Table 14 together with Figure 51 summarizes a set of measurements that was done for this test board.

<table>
<thead>
<tr>
<th>#</th>
<th>Supply voltage [V]</th>
<th>Load resistance [Ω]</th>
<th>Shielded supply cable</th>
<th>Shielded load cable</th>
<th>Extra switch node area</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>1+2+3</td>
</tr>
<tr>
<td>2</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>2+3</td>
</tr>
<tr>
<td>3</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>3</td>
</tr>
<tr>
<td>4</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>–</td>
</tr>
</tbody>
</table>

Table 14. Measurement parameters for the traces in Figure 51
Figure 51. Radiated emissions measurements according to Table 14

Again we see a clear relationship between the measured radiated emissions and the switch node area. Comparing the traces in Figure 51 to the corresponding traces in Figure 46 we see that despite a slight shift in frequency they all are more or less identical, thus in this case we cannot prove any obvious effect of the distance between the top layer and the ground plane.

Since the physical difference between TB1 and TB2 is small and the radiated emission measurements coincide so well this instead suggests quite good repeatability in the measurement setup. Even though not obvious here, PCB stack-up can still be critical for the EMC performance of a board [30].
5.2.3 Test Board 3

Test board 3, TB3 is identical to TB2 except this time the ground fill on the top layer is not present, as can be seen in Appendix G. A bare PCB for TB3 is depicted in Figure 52.

![Test board 3](image)

Figure 52. Test board 3

Table 15 lists the parameters used for the measurements shown in Figure 53.

<table>
<thead>
<tr>
<th>#</th>
<th>Supply voltage [V]</th>
<th>Load resistance [Ω]</th>
<th>Shielded supply cable</th>
<th>Shielded load cable</th>
<th>Extra switch node area</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>1+2+3</td>
</tr>
<tr>
<td>2</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>2+3</td>
</tr>
<tr>
<td>3</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>3</td>
</tr>
<tr>
<td>4</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>–</td>
</tr>
</tbody>
</table>

Table 15. Measurement parameters for the traces in Figure 53
Looking at the traces in Figure 53 we see for the third time how the peaks in the measured radiated emission spectrum drop as the switch node area decreases. Carefully comparing the traces in Figure 53 with the traces in Figure 51 we see that the peaks for TB3 seem to be slightly higher than the corresponding peaks for TB2. The difference though, is quite small so I would be a little careful to draw any definite conclusions from this. On the other hand since the increase in measured radiation seem to be present in as good as all of the peaks in all of the traces this may indicate that the absence of ground fill on TB3 has indeed caused the measured radiation to increase by a small amount. So in the hunt for fractions of, or single decibels, flooding the top layer with ground copper might be worth a try.

5.2.4 Test Board 4

In engineering, optimization and trade-offs plays an important role. Test board 4, TB4 was designed with such an issue in mind. By now it should be pretty apparent that minimizing the switch node area in a buck converter design is one the objectives if EMC performance is of any concern.

Even if the efficiency of a buck converter is typically high, thermal design considerations can often not be neglected. This often means adding copper areas for cooling purposes. If such a copper area is needed at the switch node for thermal reasons, we obviously have a typical trade-off to consider.
To see if we somehow can get away with a larger copper area at the switch node, TB4 was designed with the larger part of the switch node placed on an inner layer with ground on both adjacent layers, see Appendix H for details.

An assembled PCB for TB4 can be seen in Figure 54. Since the switch node is buried on an inner layer, modification of the area is not possible.

![Figure 54. Test board 4](image)

The traces in Figure 55 show the results of the radiated emission measurements according to Table 16.
Table 16. Measurement parameters for the traces in Figure 55

<table>
<thead>
<tr>
<th>#</th>
<th>TB#</th>
<th>Supply voltage [V]</th>
<th>Load resistance [Ω]</th>
<th>Shielded supply cable</th>
<th>Shielded load cable</th>
<th>Extra switch node area</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>4</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>N/A</td>
</tr>
<tr>
<td>2</td>
<td>1</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>–</td>
</tr>
<tr>
<td>3</td>
<td>2</td>
<td>23.5</td>
<td>5</td>
<td>No</td>
<td>No</td>
<td>–</td>
</tr>
</tbody>
</table>

Turning our attention to Figure 55 we notice that the peak magnitudes are very similar between all the different traces. This means that the extra switch node area on the inner layer of TB4 does not affect the radiated emissions in any noticeable way. Looking back at the measurements for TB1 or TB2 we can see that by placing the switch node on an inner layer and surrounding it with ground planes, the measured radiated emission was made to drop by more than 10 dB.

The cooling effect for a copper area on an inner layer is of course not as good as for a matching area on an outer layer. Nevertheless, as these measurements indicate, a buried copper area can be a very good compromise. If the copper area is to be used as a heat sink it is also important to have a good thermal connection to the heat source, this is typically achieved through multiple vias.
5.2.5 Test Board 5

Up until now all the test boards have been very similar in design, this to help us to as accurate as possible investigate the impact of single layout parameters, such as switch node area or distance to ground plane. At the same time this also allowed us to compare the results between all the different test boards without introducing too many unknowns or uncertainties. The design of the fifth and last test board described in this text differs a little more from the rest of the test boards previously discussed. The purpose with test board 5, TB5 is to investigate the significance of the switch loop area when it comes to radiated emission measurements. Details of the PCB design for TB5 are found in Appendix I, and a picture of an unpopulated PCB is presented in Figure 56.

![Figure 56. Test board 5](image)

Because we want a controlled current path for the switch current, this board was routed without any ground planes. Also remember that even an isolated conducting plane in proximity of a loop will affect the electrical properties of the loop, see for example [26]. Furthermore the switch loop was made extendable by the areas marked 1 & 2 in Figure 56. In order to not simultaneously alter the switch node area this loop extension was physically inserted at the ground end of the switch loop.
Refer to section 3.3 and Figure 14 for a description of the switch loop. If we take a look at the traces in Figure 57 we can see the result of the measurements according to Table 17.

![Graph showing radiated emissions measurements](image)

**Figure 57. Radiated emissions measurements according to Table 17**

<table>
<thead>
<tr>
<th>#</th>
<th>Supply voltage [V]</th>
<th>Load resistance [Ω]</th>
<th>Shielded supply cable</th>
<th>Shielded load cable</th>
<th>Extra switch loop area</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>23.5</td>
<td>2.35</td>
<td>No</td>
<td>No</td>
<td>–</td>
</tr>
<tr>
<td>2</td>
<td>23.5</td>
<td>2.35</td>
<td>No</td>
<td>No</td>
<td>1</td>
</tr>
<tr>
<td>3</td>
<td>23.5</td>
<td>2.35</td>
<td>No</td>
<td>No</td>
<td>1+2</td>
</tr>
</tbody>
</table>

**Table 17. Measurement parameters for the traces in Figure 57**

While adding the first extra loop area changed the result a little bit, addition of the second loop area led to a substantial increase in measured radiated emissions. In search for some understanding of how this increase in loop area raised the measured emissions by well over 20 dB for all of the peaks in Figure 57, a new set of measurements according to Table 18 was made.
As can be seen from the results plotted in Figure 58, the previous distinct increase in radiated emissions with the addition of loop area 1+2 is now absent.

Since the only difference is that the cables were now shielded, we conclude that the increase in measured radiated emission is dominated by signals emanating from the cables. Again looking at the traces in Figure 58 we also see how the general levels of emission are fairly similar between all three traces. From this fact we conclude that increasing the switch loop area does not seem to substantially increase the radiated emissions in terms of emissions related to the current in the loop itself.

Also note that $T\# [1]$ and $T\# [2]$ in Figure 58 are very similar to $T\# [1]$ and $T\# [2]$ respectively in Figure 57. The fact still remains though: that with unshielded cables the loop area clearly affected the measured radiated emission levels. To see whether one or both of the cables are the cause for the increased measured radiation, two measurements according to Table 19 were performed.

<table>
<thead>
<tr>
<th>#</th>
<th>Supply voltage [V]</th>
<th>Load resistance [Ω]</th>
<th>Shielded supply cable</th>
<th>Shielded load cable</th>
<th>Extra switch loop area</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
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<td>2.35</td>
<td>Yes</td>
<td>Yes</td>
<td>–</td>
</tr>
<tr>
<td>2</td>
<td>23.5</td>
<td>2.35</td>
<td>Yes</td>
<td>Yes</td>
<td>1</td>
</tr>
<tr>
<td>3</td>
<td>23.5</td>
<td>2.35</td>
<td>Yes</td>
<td>Yes</td>
<td>1+2</td>
</tr>
</tbody>
</table>

Table 18. Measurement parameters for the traces in Figure 58

Figure 58. Radiated emissions measurements according to Table 18
<table>
<thead>
<tr>
<th>#</th>
<th>Supply voltage [V]</th>
<th>Load resistance [Ω]</th>
<th>Shielded supply cable</th>
<th>Shielded load cable</th>
<th>Extra switch loop area</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>23.5</td>
<td>2.35</td>
<td>No</td>
<td>Yes</td>
<td>1+2</td>
</tr>
<tr>
<td>2</td>
<td>23.5</td>
<td>2.35</td>
<td>Yes</td>
<td>No</td>
<td>1+2</td>
</tr>
</tbody>
</table>

Table 19. Measurement parameters for the traces in Figure 59

Figure 59 shows the result from those measurements, and here we can see that in both cases the measured radiation levels are lower compared to when none of the cables were shielded but still higher compared to when both of the cables were shielded. See T# [3] in Figure 57 that T# [3] in Figure 58 for those cases respectively.

![Figure 59. Radiated emissions measurements according to Table 19](image)

This means that both the supply cable and load cable contribute to the radiated emissions. From Figure 59 we can also deduce that in this case slightly more radiated emissions seem to originate from the load cable compared to the supply cable. Reviewing Figure 57 and Figure 58 we can also see that except for the obvious differences in peak amplitudes, changing the switch loop area also seem to change the spectral content of the radiated emissions. With this hint in mind we now take another look at the voltage waveform at the switch node. Figure 60 shows an oscilloscope measurement of the waveform at the switch node of TB5 with parameters according to #3 in Table 17.
While the oscilloscope trace for TB1 resembles a decent rectangle pulse, the same pulse for TB5 with the extra loop area suffers from severe over and under shoots.
with associated ringing. From this distorted waveform we can easily understand a change in spectral content. With aid of the waveform in Figure 60 we now conclude that the extra inductance associated with the increase of the switch loop area for TB5, alters the appearance of the switch node voltage, and this at the same time causes noise to propagate out onto attached cables. This noise on the cables then substantially contributes to the overall measured radiated emission levels.

Therefore, even if voltage is the main driving force behind the radiated emissions, current and current loops in a buck converter should not be neglected. The above described measurements for TB5 also serve as a good example of how complex and multifaceted an EMC issue can often be, and as a consequence, the importance of careful consideration when it comes to conclusions about cause and effect. Even though a little off topic for this text we can also note that the layout for TB5 would in a general case be a rather poor layout, not only from an EMC point of view but also from an electrical performance standpoint, yet another reason to not lessen the importance of the current switch loop in a buck converter.
6 Summary

We have showed both theoretically–by the use of simple models and 3D electromagnetic computer simulations–and practically–by measurements on several test boards–that when measured according to CISPR 25 and for frequencies up to a few MHz, the radiated emissions from a buck converter are dominated by voltage driven mechanisms. As we also have seen this means that in a PCB layout, all areas which are connected to a time varying voltage are potential sources of radiated emissions. As a consequence, to keep radiated emissions at a minimum, such areas should be kept as small as possible and or electrically screened.

Once again, do not let yourself to believe that currents are of no importance when it comes to radiated emission measurements, one reason of course: that when it comes to time varying fields there is an intimate relation between electric and magnetic fields. At last, if you read between the lines in this text I am sure you can draw conclusions I have yet to discover and probably ask yourself even more questions that hopefully will inspire you to further investigation in the matter!
7 References


[29] Texas Instruments Incorporated, 3.5V to 60V Input, 2.5A, Step Down Converter with Eco-Mode™, 2010, SLVSA86A.

Appendix A – Measurement Setup

The measurements chamber is an ETS-Lindgren type S81 shielded enclosure with dimensions: $3 \times 5 \times 2.3 \text{ m}$. Notice that this chamber is not equipped with any kind of electromagnetic energy absorbers. Inside the chamber a $1.5 \times 3.2 \text{ m}$ big and $0.9 \text{ m}$ high metal table serves as test table and local ground plane. The power to the device under test is supplied through two LISNs from Fischer Custom Communications, Inc. with part no: FCC-LISN-5-32-1-01-CISPR25 (measurement ports terminated with 50 Ω). The measurement antenna is a battery fed active 1 m long vertical monopole. The output from the antenna is connected to a 9 kHz–3.6 GHz N9010A EXA Signal Analyzer from Agilent Technologies. The test measurement sweeps are controlled from a computer program for RF emission tests named REMI® (Radio Frequency emission), release 2.134.
Appendix B – Buck Converter Schematic
## Appendix C – Bill of Materials

<table>
<thead>
<tr>
<th>Count</th>
<th>RefDes</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>U1</td>
<td>TPS54260DGQ</td>
<td>IC, Step-Down Converter</td>
</tr>
<tr>
<td>1</td>
<td>C11</td>
<td>4.7nF</td>
<td>Capacitor, Ceramic, 1206, 630V, X7R, 10%</td>
</tr>
<tr>
<td>3</td>
<td>C2, C4, C5</td>
<td>4.7µF</td>
<td>Capacitor, Ceramic, 1210, 50V, X7R, 10%</td>
</tr>
<tr>
<td>1</td>
<td>C12</td>
<td>10nF</td>
<td>Capacitor, Ceramic, 0603, 50V, X7R, 10%</td>
</tr>
<tr>
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<td>C7</td>
<td>4.7nF</td>
<td>Capacitor, Ceramic, 0603, 50V, X7R, 10%</td>
</tr>
<tr>
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<td>C6</td>
<td>100nF</td>
<td>Capacitor, Ceramic, 0603, 50V, X7R, 10%</td>
</tr>
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<td>2</td>
<td>C8, C9</td>
<td>47µF</td>
<td>Capacitor, Ceramic, 1210, 10V, X7R, 10%</td>
</tr>
<tr>
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<td>C1, C3</td>
<td>220µF</td>
<td>Capacitor, Aluminium Electrolytic, Case G, 50V, 20%</td>
</tr>
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<td>B360B-13-F</td>
<td>Diode, Schottky, 60V, 3A, SMB</td>
</tr>
<tr>
<td>1</td>
<td>D2</td>
<td>S2M</td>
<td>Diode, 1000V, 2A, SMB</td>
</tr>
<tr>
<td>2</td>
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<td>Inductor, SMT, 4.32A, 26mΩ, MSS1048-103MLB</td>
</tr>
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<td>R6</td>
<td>330k</td>
<td>Resistor, Chip, 1/16W, 1%</td>
</tr>
<tr>
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<td>R1</td>
<td>390k</td>
<td>Resistor, Chip, 1/16W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R5</td>
<td>82k</td>
<td>Resistor, Chip, 1/16W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R3</td>
<td>20k</td>
<td>Resistor, Chip, 1/16W, 1%</td>
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<td>62k</td>
<td>Resistor, Chip, 1/16W, 1%</td>
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<tr>
<td>1</td>
<td>R4</td>
<td>20k</td>
<td>Resistor, Chip, 1/16W, 1%</td>
</tr>
<tr>
<td>2</td>
<td>J1, J2</td>
<td>MKDS 1/2-3,5</td>
<td>Terminal Block, 2 pos, 10A, 3.5mm pitch</td>
</tr>
</tbody>
</table>
Appendix D – Assembly Drawing
Appendix E – Test Board 1
Appendix G – Test Board 3
Appendix H – Test Board 4
Appendix I – Test Board 5